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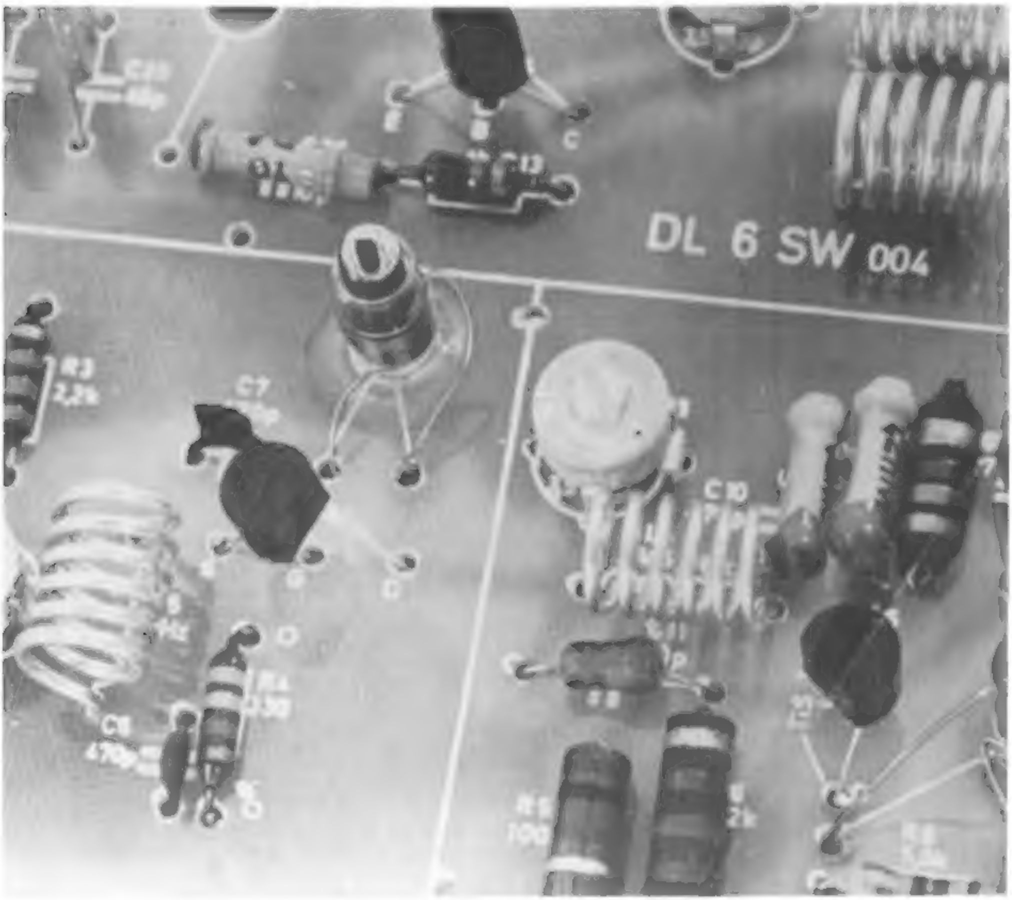
VHF COMMUNICATIONS

A PUBLICATION FOR THE RADIO AMATEUR
ESPECIALLY COVERING VHF, UHF AND MICROWAVES

VOLUME NO. 1 FEBRUARY 1969 US $ 1.00



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VHF COMMUNICATIONS, the international edition of the well-established German publication UKW-BERICHTE, is a quarterly amateur radio magazine especially catering for the VHF/UHF/SHF technology. Published in February, May, August and November.

The subscription price is US $ 3.00 or national equivalent per year. Individual copies are available at $ 1.00, or equivalent, each.

Subscriptions, orders of individual copies, purchase of printed circuit boards and advertised special components, advertisements and contributions to the magazine should be addressed to the national representatives. It is important that any change of address be reported as soon as possible to ensure the correct and punctual arrival of the publication. Please give your address in block letters and make sure to place sufficient postage.

Publisher: VERLAG UKW-Berichte, Hans J. Dohlus, DJ 3 QC, D-8520 ERLANGEN, Gleiwitzer Str. 45, Federal Republic of Germany, Tel. (0 91 31) 3 33 23

Editors: Robert E. Lentz, DL 3 WR, Terry D. Bittan, G 3 JVQ, DJ Ø BQ

Printed in the Federal Republic of Germany by Richard Reichenbach KG, D-8500 Nuernberg, Krelingstraße 39.

VERTRETUNGEN: REPRESENTATIVES:

Austria	Hans J. Dohlus, DJ 3 QC, D-8520 ERLANGEN, Gleiwitzer Straße 45, see Germany
Australia	WIA, PO Box 67, East Melbourne 3002, Victoria
Belgium	E. Drieghe, ON 5 JK, B-9160 HAMME, Kapellestr. 10, Telefon (052) 49457, PCR 588111
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France	Christiane Michel, F 5 SM, F-89 PARLY, les Pillés, CCP PARIS 16.219-66
Finland	Timo Ekko, OH 1 SM, PORI, Kalevanpuisto 63 C, Postgiro 47-10 848
Germany	Verlag UKW-BERICHTE H. Dohlus oHG, D-8520 ERLANGEN, Gleiwitzer Str. 45, Telefon (09131) 3 33 23 + 6 33 88, Deutsche Bank Erlangen Kto. 476 325, Postscheck 304 55 Nürnberg
Holland	S. Hoogstraal, PA ø MSH, ALMELO, Oranjestraat 40, giro 137 2 282, telefon (05490) — 1 26 87
Italy	STE s.r.l. (I 1 GM) via maniago 15, I-20134 MILANO, Tel. 21 78 91, Conto Corrente Postale 3/44968
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Switzerland	Hans J. Dohlus, DJ 3 QC, D-8520 ERLANGEN, Gleiwitzer Straße 45, see Germany
United Kingdom	MICROWAVE MODULES Ltd., 4 Newling Way, WORTHING/SSX. Tel. 0903-64301
USA-East Coast	VHF COMMUNICATIONS Russ Pillsbury, K 2 TXB, & Gary Anderson, W 2 UCZ, 915 North Main St. JAMESTOWN, NY 14701
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USA-West Coast	Darrel Thorpe, Circuit Specialists Co. Box 3047, SCOTTSDALE AZ 85257, Tel. 602-945-5437

VOLUME 1 FEBRUARY 1969 EDITION 1

PUBLISHER: VERLAG UKW-BERICHTE
Hans J. Dohlus, DJ 3 QC
Gleiwitzer Strasse 45
Fed. Republic of Germany

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VHF COMMUNICATIONS

The first edition of VHF COMMUNICATIONS represents the birth of a new amateur radio magazine. However, VHF COMMUNICATIONS has not "appeared out of the blue" but is the international edition of the well known German language publication UKW-BERICHTE which has regularly appeared since 1960. The increased editorial team intends to forward the same ideals in both VHF COMMUNICATIONS and UKW-BERICHTE: To publish a magazine from radio amateurs for radio amateurs, specializing itself to the VHF, UHF and microwave technology.

The principle task of VHF COMMUNICATIONS is to provide exact and extensive assembly instructions for transmitters, receivers, converters, complete tranceivers, antennas, measuring and auxiliary equipment etc., allowing them to be easily duplicated. The great advantage for our readers is that all the epoxy printed circuit boards and other specialized components, such as trimmers, coil formers, complicated metal-work, etc., are available from our representatives at moderate prices.

The ever increasing demands made on the efficiency and performance of amateur radio equipments as well as the continuously advancing technology resulting from this, makes it more and more difficult for the individual amateur to assemble his own equipment. Even though the turnover of amateur radio equipment manufacturers is steadily increasing, we feel that the majority of amateurs would prefer to build their own high-performance equipment if only the corresponding instructions and assistance were available. This is where we would like to help by providing detailed descriptions and supplying the special components. We believe that these ideals have been responsible for the success of the German language edition. We hope that our readers throughout the world will agree with us and help us to forward our ideals and the cause of amateur radio.

Vy 73, the publishers

H. Dohlus R. Lentz Terry Bittan

DJ 3 QC DL 3 WR G 3 JVQ / DJ ∅ BQ

A 2 METRE CONVERTER WITH FIELD EFFECT TRANSISTORS

from W. von Schimmelmann, DL 6 SW

1. INTRODUCTION

A great deal has been written recently regarding the excellent qualities of field effect transistors with regard to cross-modulation and intermodulation rejection. Since field effect transistors are now available on the market for roughly the same price as a good electronic tube, the decision was made to examine the RF characteristics in experiments with a 2 metre converter. The task was to replace a tubed converter, if the cross-modulation and amplification were in the same order as that obtained with tube-equipped types.

This target was met, as can be seen in Section 5, with the following converter. Several different tubed and transistorized converters were measured under the same conditions and are given as a comparison.

TIS 34 junction field effect transistors manufactured by the firm of Texas Instruments were used in this design and proved to be excellent and insensitive to overloading.

The first attempt was built-up using a single RF amplifier stage, the subsequent mixer being fed via an RF transformer. The pre-amplifier and mixer stages were each equipped with a TIS 34 transistor, whereas the oscillator and tripler stage were equipped with two BF 224 transistors. Due to the fact, however, that the transconductance of the junction field effect transistors available at present is extremely low, the amplification proved to be so insufficient that the noise figure of the mixer was noticeably adversely effecting the total noise figure. The attempt was then made to increase the amplification with the aid of a high LC ratio at the RF transformer as well as to reduce the loss by dispensing with alignment cores. This, however, only brought a sensitivity figure of approximately 3.5, whereby the adjustment of the neutralization was found to be so critical that it could be only carried out exactly with the aid of a sweep measuring set.

The following circuit was built-up using two RF amplifier stages so that the amplification is sufficient to allow the second RF stage and the mixer to be fed at low impedance. This meant that the neutralization was no longer critical and that the field effect transistors obtained the source impedance of approx. 1 kΩ needed for the lowest noise figure.

2. CIRCUIT DESCRIPTION

The circuit as shown in Fig. 1 comprises two RF amplifier stages, the mixer and a two stage oscillator circuit. The oscillator is equipped with a 38.666 MHz crystal and is followed by a tripler stage multiplying the crystal frequency to 116 MHz.

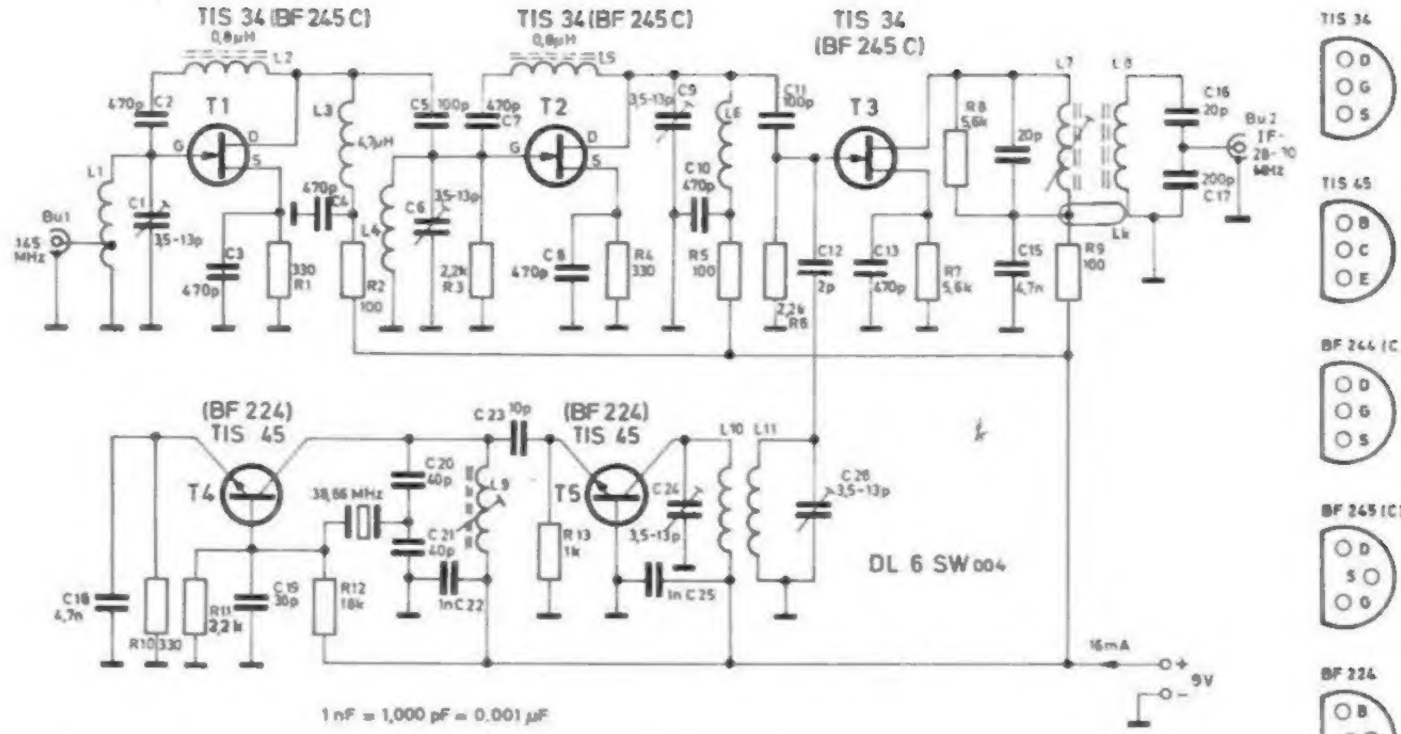


Fig. 1 Circuit Diagram of the 2 Metre Field Effect Transistor Converter

Junction field effect transistors of the type TIS 34 are used in the two RF stages and for the mixer (in contrast to metal oxide silicon field effect transistors MOS FET). The TIS 34 is a depletion-mode FET.

Both RF amplifier stages operate as a neutralized source circuit, whereby the inductances L 2 and L 5 form, together with the feedback capacity of the transistor, a neutralizing circuit with resonance at 145 MHz. The capacitors C 2 and C 7 serve merely as DC blocking.

The input circuit of the converter comprising L 1 and C 1 possesses a low-impedance tapping for the connection of the antenna. An intermediate link consisting of L 4 and C 6 is to be found between the two RF amplifier stages. The drain voltage is fed to transistor T 1 in parallel to the resonant circuit,via chocke L 3.

The RF signal is further amplified in transistor T 2 and is fed via a second intermediate link consisting of L 6 and C 9 to the mixer, which does not require neutralizing. When the oscillator signal is absent, a voltage drop of approximately 1.2 V appears across the source resistor R 7 of the mixer transistor T 3 which causes a quiescent current of approximately 0.2 mA. If the voltage drop across R 7 reaches approx. 2.1 V, the amplitude of the 116 MHz oscillator will be sufficient. The drain current will thereby be increased to approximately 0.35 mA.

The primary of the IF transformer following the mixer is damped with resistor R 8. A coupling link is to be found at the cold end of inductances L 7 and L 8. After connection of the subsequent receiver via the capacitive voltage divider comprising C 16 and C 17, a bandwidth of roughly 2.5 MHz is obtained with a dip of approx. 3 dB at 29 MHz.

The oscillator and tripler stage are equipped with "normal" bipolar transistors type BF 224, which are silicon planar NPN transistors in epoxy casings. They can of course be replaced with similar transistor types such as 2 N 918, 2 N 914 (BSY 21), BF 115, BC 108 or BC 130.

The oscillator transistor T 4 operates with a 38.666 MHz overtone crystal in a circuit previously described in (1) and (2). The feedback is achieved using a capacitive voltage divider. The common-base tripler transistor T 5 is loosely coupled via C 23 to the hot end of the oscillator tank circuit. The 116 MHz transformer at the collector of transistor T 5 is inductively coupled without the need of a coupling link since the inductance L 10 and L 11 are so arranged that they are directly adjacent to another. The 116 MHz signal is then finally fed via C 12 to the gate of mixer transistor T 3.

Since the current consumption of the whole converter is only approximately 16 mA at an operating voltage of 9 V, it was decided to dispense with an AC power supply. Instead of this, six 1.5 V dry batteries were placed in series. The resulting operating period is roughly 200 to 300 hours. This form of power supply has the additional advantage of offering an exceptional frequency stability.

3. MECHANICAL ASSEMBLY (Fig. 2 and 3)

The converter is built-up on a printed circuit board with the dimensions of 63 mm by 113 mm (see Fig. 2a). The printed circuit board is divided into four different chambers by brass or copper strips. One chamber is used for the antenna input circuit with transistor T 1, a second for the resonant circuit comprising L 4, C 6 and the transistor T 2, and a third containing the mixer stage and the IF transformer. The elongated fourth chamber accomodates the crystal oscillator and the tripler stage. The coupling link at the cold end of the IF transformer L 7/L 8 (see Fig. 1) represents loop in the form of an elongated "0", each former having only a half turn. The link is made up from 0.2 mm (32 AWG) diameter wire - a thicker diameter wire would cause the coupling to be too great.

The metal casing of the crystal should be grounded to the nearest ground point of the printed circuit board.

It is strongly advised that the converter be enclosed in a metal cabinet equipped with coaxial connectors. If the batteries are not contained within the cabinet, the operating voltage should be fed via feed-through capacitors. Unwanted short wave signals have been observed to be induced into the converter in its unscreened state.

With the aid of the circuit diagram (Fig. 1), the parts list (Section 3.1) and the location plan (Fig. 2b), it is possible to commence the mounting of the components on to the printed circuit board. This is done in the following order: The resistors are the first components to be mounted which are then followed by the capacitors. The transistors are then mounted after the inductances. The transistor connections, as seen from below, are given in Fig. 1. Finally, the crystal and the screening plates, which form the individual chambers, are mounted. The completed converter is shown in Fig. 3.

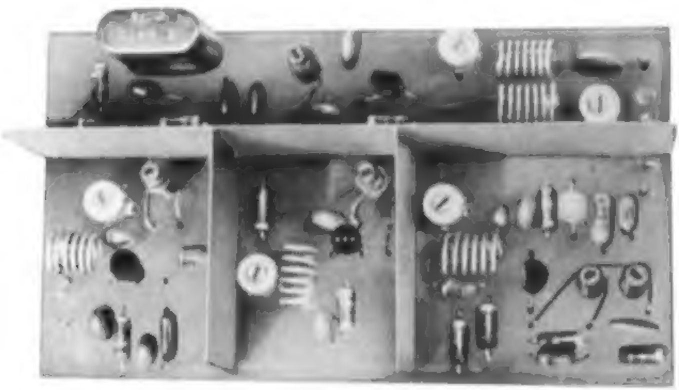


Fig. 3 The Assembled 2 Metre FET Transistor Converter

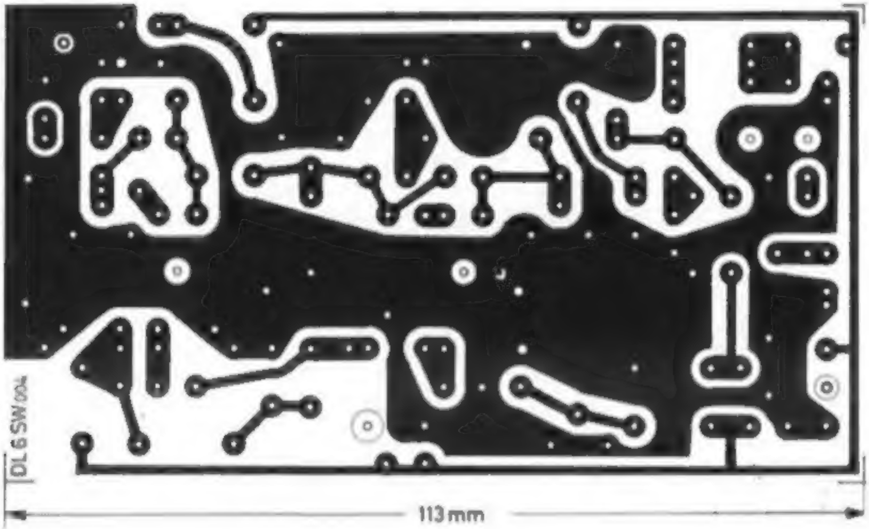


Fig. 2a Conducting Side of the Printed Circuit Board
DL 6 SW Field Effect Transistor Converter for 2 Metres

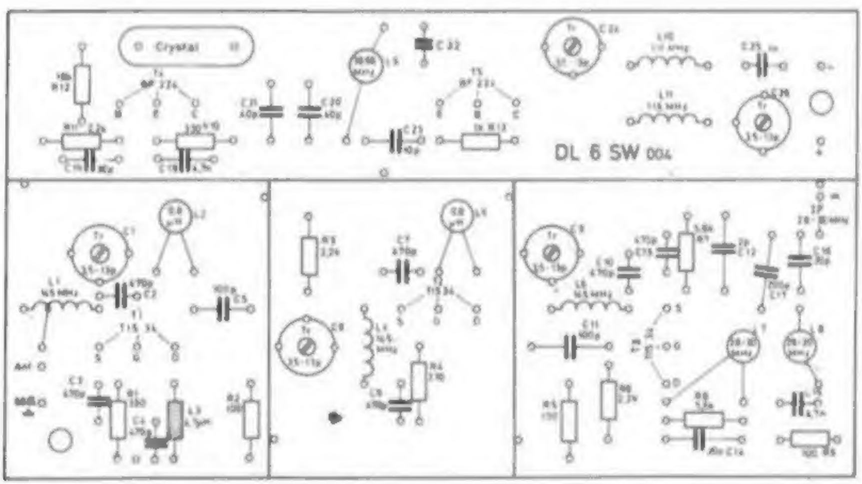


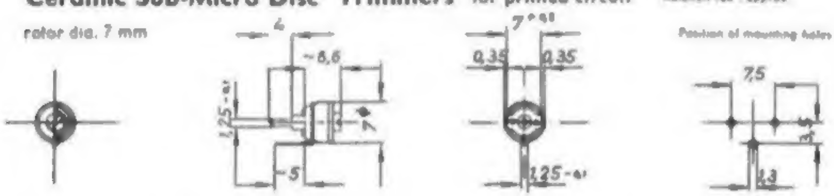
Fig. 2b Location Plan

3. 1. COMPONENTS

R 1, R 4 and R 10:	330 Ω;	R 2, R 5 and R 9:	100 Ω	All resistors are
R 3, R 6 and R 11:	2. 2 kΩ;	R 7 and R 8:	5. 6 kΩ	rated at 0. 3 watt.
R 12:	18 kΩ;	R 13:	1 kΩ	

C 1, C 6, C 9, C 24 and C 26: 3. 5 to 13 pF ceramic trimmers
C 2, C 3, C 4, C 7, C 8, C 10 and C 13: 470 pF ceramic disc capacitors
C 15 and C 18: 0. 0047 μF ceramic disc capacitor
C 22 and C 25: 0. 001 μF ceramic disc capacitor
C 12: 2 pF ceramic disc capacitor
C 5 and C 11: 100 pF ceramic tubular capacitor
C 14 and C 16: 20 pF ceramic tubular capacitor
C 17: 200 pF ceramic tubular capacitor
C 19: 30 pF ceramic tubular capacitor
C 20 and C 21: 40 pF ceramic tubular capacitor

Ceramic Sub-Micro Disc Trimmers for printed circuit



T 1, T 2 and T 3: TIS 34 (BF 245) from Texas Instruments
T 4 and T 5: BF 224 from Texas Instruments
Q Overtone crystal 38. 666 MHz HC 6/U

3. 2. COIL DATA

L 1, L 4 and L 6: 6 turns of 1 mm (18 AWG) dia. silver-plated copper wire wound on a 5 mm dia. former, self-supporting and with a length of 10 mm. The tap of L 1 is made 1 turn from the cold end.

L 2 and L 5: 10 turns of 0. 3 mm (29 AWG) dia. enamelled silk-covered wire on a 3. 5 mm former, 0. 8 μH, VHF core.

L 7 and L 8: 20 turns of 0. 3 mm (29 AWG) dia. enamelled silk-covered wire on a former and core as given for L 2 and L 5.

L 9: 15 turns of 0. 3 mm (29 AWG) dia. enamelled silk-covered wire on a former and core as given for L 2 and L 5.

L 10 and L 11: 7 turns of 1 mm (18 AWG) dia. silver-plated copper wire wound on a 5 mm dia. former, self-supporting and with a coil length of 10 mm.

L 3: 4. 7 μH RF choke or 1 metre of 0. 2 mm (32 AWG) dia. enamelled copper wire on a 4 mm dia. former.

4. ALIGNMENT

The neutralizing coils L 2 and L 3 are aligned to 145 MHz before mounting with the aid of a grid-dip meter and a parallel capacitor of 1. 5 pF, whereby care should be take that the connecting leads of the coils are not longer than under mounted conditions. If an inductance meter is available, the coils can directly aligned to an inductance of 0. 8 μH. The other resonant circuits can be coarsely aligned to the relative frequency with the aid of a grid-dip meter, in their mounted state.

The crystal oscillator only operates at its nominal frequency and does not tend to oscillate at spurious frequencies. The frequency of the crystal can be slightly pulled by adjusting the value of capacitor C 19 between 10 and 30 pF. The tripler stage is aligned for a maximum voltage drop across the source resistor R 7 of 2. 1 V. When the oscillator is inoperative, the voltage drop across resistor R 7 should read approximately 1. 2 V.

The IF transformer is now aligned for maximum noise. The 10 metre receiver should be set to 28. 3 MHz for the alignment of the primary stage comprising L 7 and C 14 and to approximately 29. 7 MHz for the alignment of the output circuit. The input stage is aligned in similar manner at 145 MHz, the intermediate link L 4 and C 6 at 145. 5 MHz and the mixer stage with L 6 and C 9 at 144. 5 MHz for maximum noise indication. This alignment must be repeated until an equal amplification is obtained over the whole range from 144 MHz to 146 MHz. The neutralization does not normally require realignment. The most simple means of aligning the converter is with the aid of a sweep measuring set, but this need not be mentioned here in great detail. The response curve of the converter can be adjusted to obtain a bandwidth of 2. 5 MHz with a dip of roughly 2 dB at 145 MHz.

5. MEASURED VALUES

The following table contains the more important performance specifications of the described 2 metre converter. The performance of some other converters were measured under the same conditions and are given here for comparison. The other converters tested were a converter equipped with the PC 88 tube, one with a PC 97 (3) (4) and the commercial transistor type MB 22 manufactured by the German company of Semcoset. The converter output signal was fed in all cases to a Drake R 4 whose bandwidth amounted to 2. 7 kHz.

Unit / Measurement	TIS 34 Converter	PC 88 Converter	PC 97 Converter	Semcoset MB 22
Noise figure	2. 0	3. 5	2. 7	2. 7
Amplification (dB)	22	33	18	27
Image rejection (dB)	65	60	60	65
Cross-modulation (mV)	35	35	40	4
Wipe-out effect (mV)	20	10	50	2
IF rejection (dB)	65	65	80	75

Notes:

The cross-modulation measurement was made by feeding an unmodulated 1 µV signal at a frequency of 144.5 MHz to the converter. A second 30 % modulated signal was then also fed to the converter and its level increased until a 1 % cross-modulation was observed. The value of the modulated input signal (at 145.5 MHz) is then given in mV.

The wipe-out effect was measured by feeding an unmodulated 2 µV signal to the converter at a frequency of 144.5 MHz. A second signal (at 145.5 MHz) was then fed to the converter and the level increased until a reduction of 3 dB was observed on the first signal. The value in mV of this second signal is given in the above table.

The excellent characteristics of the DL 6 SW FET Converter may also be utilized for satellite band reception in the frequency range of 136 to 137 MHz - i.e. for the monitoring of weather satellites. It is merely necessary to replace the 38.66 MHz crystal with one of 53.7 MHz and to align the transistor stage with T 5 to the first harmonic of the crystal frequency (107.4 MHz) as well as the two pre-amplifier stages T 1 and T 2 to the frequency range of 135.4 MHz to 137.4 MHz.

If a different intermediate frequency is to be used instead of 28 to 30 MHz, the necessary crystal frequency should be calculated without difficulty

6. REFERENCES

(1) RCA Transistor Manual, Page 440

(2) M. Kindler: Ein Transistorkonverter für 145 MHz mit guter Vorselektion
UKW-BERICHTE 6 (1966) H. 2, Pages 88 to 97

(3) R. Lentz: Der PC 97-Konverter
DL-QTC 34 (1963) H. 6, Pages 268 to 273

(4) Ch. Junge: Der PC 900-Konverter
DL-QTC 35 (1964) H. 8, Pages 470 and 471

EDITORIAL NOTES

The values given in Section 5 cannot be considered as absolute values because these measurements are - except for the amplification measurement - extremely critical and difficult to reproduce. However the comparison can be clearly seen and we consider the table to be informative.

In the mean time, a great number of the described converters have been built, measured and tested under practical conditions. This has shown that the given specifications have been achieved and, in some cases, exceeded. The alignment is uncritical and can be achieved without difficulties. The practical test was carried out by DL 3 SPA at a good clear location at an altitude of 600 metres (1815 ft.) using a 16 element antenna. The test proved that the converter is, in all respects, equal to a good tubed type and, in contrast to past converters equipped with conventional transistors, showed that the cross-modulation characteristics were not worse than those of a good tube converter.

We would like to point out that european amaterus can use the FET transistor types BF 244 and BF 245 instead of the TIS 34. They are electrically completely identical to the TIS 34 (socket connections are given in Fig. 1). If the transistor TIS 45 is used for the two oscillator transistors T 4 and T 5 a greater auxiliary frequency level is obtained which enables the amplification of the converter to be increased to 25 to 28 dB.

For european readers the printed circuit boards, coil formers and trimmers are available from:

Verlag UKW-BERICHTE, Hans J. Dohlus, DJ 3 QC, D 8520 Erlangen, Gleiwitzer Str. 45, Postscheck-Konto 30455 Nürnberg, West Germany (see classified advertisement).

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PHASE-LOCKED OSCILLATOR FOR TRANSMIT AND RECEIVE MIXERS AMATEUR RADIO EQUIPMENT

by K. P. Timmann, DJ 9 ZR and V. Thun, DJ 7 ZV

1. INTRODUCTION

On observing newly developed professional VHF transmitters and receivers, it will be noticed that the spurious signal and cross-modulation characteristics have been improved. For instance, equipment which has been especially developed for aeronautical radio communications is often equipped with field-effect-transistors in the input and first mixer stages. This is followed by narrow band filters with steep skirts and high attenuation far from the centre frequency, e.g. crystal filters. It is possible in this manner to obtain an extremely low cross-modulation in the input stages (virtually square dynamic characteristic of the field-effect-transistors) and to ensure that the frequencies involved in the cross-modulation do not reach the IF stages.

The necessary frequency stability and low spurious signal level of the RF voltage supplied from the auxiliary oscillator to the mixer can be obtained using the phase-locked oscillator principle. The following paragraphs briefly describe three methods of generating the auxiliary frequency as well as mentioning the advantages and disadvantages of each. This is followed by a detailed description of a practical phase-locked oscillator circuit.

2. ADVANTAGES AND DISADVANTAGES OF THE VARIOUS OSCILLATOR TYPES

2.1. FREE-RUNNING OSCILLATORS

Various attempts have been made recently to design free-running oscillators with high frequency stability at frequencies over 50 MHz. Such oscillators possess low spurious signal components, i.e. the spurious signals and harmonics are far lower than the required signal which means that they will not cause any adverse effect after frequency multiplication. The author has built up several such oscillator circuits, some comprising ceramic resonant circuit components and silicon transistors with an output capacitance of less than 0.2 pF. The frequency stability obtained with these oscillators was however not satisfactory because it was not sufficient for single sideband operation on the VHF bands.

2.2. OSCILLATORS OPERATING ACCORDING TO THE FREQUENCY SYNTHESIS (SUPERHETERODYNE) PRINCIPLE

In this principle, the frequency of a self-excited oscillator is mixed with the fixed frequency of a crystal controlled oscillator in order to obtain the variable sum frequency. Since it is only possible to build-up sufficiently stable free-running oscillators up to a frequency of approximately 10 MHz without extensive means, it is necessary to use a certain amount of filtering and a careful construction to ensure that the spurious signals do not cause images, whistle or noise points in the receiver.

2.3 OSCILLATORS OPERATING ACCORDING TO THE PHASE-LOCKED PRINCIPLE (OSCILLATOR CONTROLLED BY MEANS OF FREQUENCY OR PHASE DISCRIMINATION).

In this principle, the frequency of a self-excited oscillator is controlled if its frequency deviates from the nominal value. This is carried out, for instance, with the aid of a motor driven capacitor or varactor diodes.

The control signal is obtained via a circuit which has the task of mixing the frequency or phase of a signal voltage obtained from the local oscillator down to approximately the frequency of the VFO (reverse of the superheterodyne principle). The reduced frequency of the local oscillator is now compared with the VFO frequency in a discriminator. The control voltage at the output of this stage is filtered and controls the frequency of the oscillator with the aid of a varactor diode.

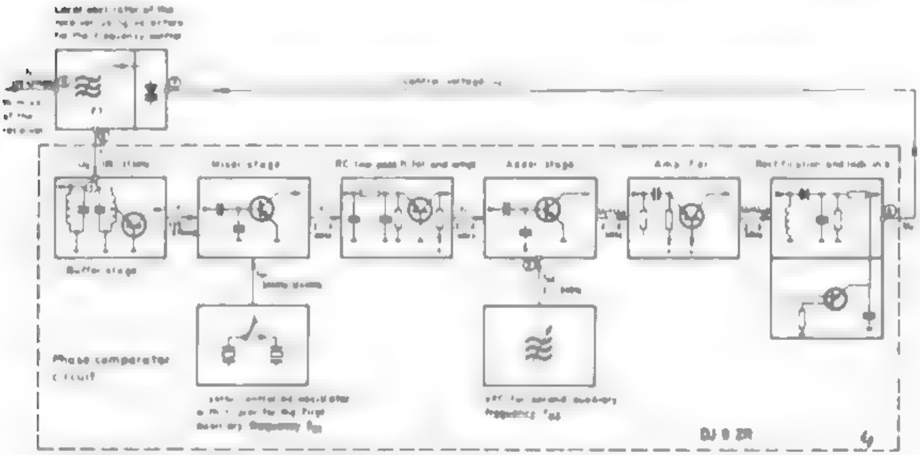


Fig 1 Block diagram of the Phase-locked Oscillator

The advantage of this method is that the auxiliary oscillator operates completely spurious-free and that the control device does not require a large amount of filtering and screening. One disadvantage of this method is that a good compromise must be carefully determined between the low-noise level and the lock-in period on varying the oscillator frequency.

Furthermore, if unfavourable RC filter links were selected for the control voltage, an oscillation of the control circuit could be observed, which can however, be easily counteracted.

3. THEORY OF OPERATION (Fig. 1)

We would like to explain the operation of the phase-locked principle with the aid of the block diagram given in Fig. 1. Let us firstly assume that the control voltage (U_c) connection is broken. An RF signal is taken from the oscillator, which is oscillating at a frequency f_1 adjacent to the required nominal frequency, and is mixed with the fixed auxiliary frequency f_{01} (or with several fixed frequencies in the subsequent stages). This has been made in order to obtain a lower frequency f_1 which is within the frequency range of the variable auxiliary oscillator (VFO). The frequency of this oscillator is adjusted to the frequency f_{02}.

A frequency or phase comparison can now be carried out in the subsequent stage.

3.1. FREQUENCY COMPARISON

In order to obtain the required sum frequency $f_{o1} + f_{o2}$ from f_1, let us firstly assume that we have a circuit in which we can compare both frequencies to another. A control voltage should be available at the output which is both of the correct sign and proportional to the frequency difference between f_1 and f_{o2}. If the control line is now reconnected, the frequency f_1 of the local oscillator will (if the polarity is correct) be reduced by means of the varactor diode until

$$f_1' \approx f_{o2} \quad \text{and thus} \quad f_1 \approx f_{o1} + f_{o2}$$

It can be seen that a residual frequency deviation exists between the frequency f_1 of the local oscillator and the nominal frequency when using a frequency discriminator. Since such circuits are more extensive than phase comparator circuits, the latter are normally employed.

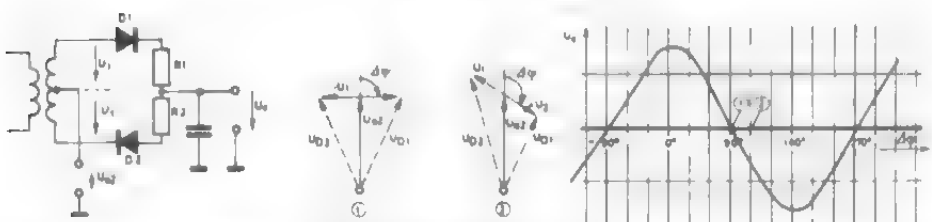
The disadvantage of the phase comparison circuit, namely that the local oscillator will be ambiguously controlled at frequency deviations of several hundred Hz between the two frequencies in question, can be compensated for with the aid of a special lock-in circuit.

3.2. PHASE COMPARISON USING A SYNCHRONOUS DEMODULATOR (Fig. 2)

Normal demodulation circuits can be used in the phase comparator circuit (phase discriminator) if the DC components are not suppressed.

The use of a synchronous demodulator (synchronous detector, product detector) as a phase discriminator is shown in Fig. 2. The diodes are connected anti-phase (Fig. 2a) and are fed in push-pull with U_1' as well as with U_{o2} in push-push. Two examples of different phase positions between U_1' and U_{o2} are shown in the vector diagram (Fig. 2b). The output voltage U_c as a function of the phase position results approximately from the difference in length of the two vectors shown as dashed lines:

$$U_c \approx \left| U_{D1} \right| - \left| U_{D2} \right|$$



a) Circuit
U_1' = IF voltage
U_{o2} = Auxiliary voltage from VFO
U_c = DC voltage resulting at the output

b) Vector Diagram at $\Delta\varphi = 90°$ and $\Delta\varphi = 120°$. The voltage U_c is proportional to the difference between the vector lengths U_{D1} and U_{D2}

c) Control voltage in dependence of the phase angle difference $\Delta\varphi$ between U_1 and U_{o2}

Fig 2 Circuit Diagram, Vector Diagram and Characteristic Curve of the Output Voltage U_c of a Synchronous Demodulator

U_c falls from a maximum value at a frequency difference of $\Delta\varphi = 0$ to zero at $\Delta\varphi = 90^0$ and will become negative at greater phase differences until $\Delta\varphi = 180^0$ is reached.

The curve differs essentially from a cosine function if U_1 and U_2 are approximately of equal amplitude. This deviation is, in our case, extremely favourable since the centre portion of the characteristic curve is flatter. (On the other hand, when using such circuits for amplitude and single sideband demodulation, the value of U_{o2} must be at least 5 times greater than U_1 to ensure a low distortion level).

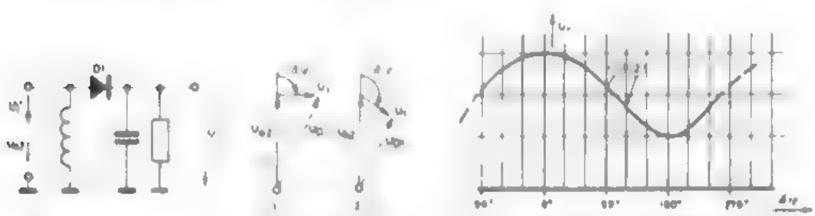
If f_{o2} and f_1' differ slightly from another, this will correspond to a continously increasing phase difference and U_c will fluctuate periodically with the difference frequency in the form of a distorted AC voltage. At greater frequency differences, the amplitude of this alternating voltage will be reduced rapidly due to the RC filter link required by the circuit.

If the control line is connected and the filtering is not too great, larger phase differences do not usualy occur which could cause the phase-lock of the local oscillator to be unsynchronized, even if the VFO is tuned or the resonant circuit of the local oscillator detuned. A phase difference between U_1 and U_{o2} slightly deviating from 90^0 is always present.

1.3. PHASE COMPARISON USING AN ENVELOPE DEMODULATOR (Fig. 3)

A simple diode circuit can often be used as a phase discriminator. Fig. 3 shows an example of such a circuit with two vector diagrams and an illustration showing the output voltage as a function of the phase difference. The output voltage U_c is approximately proportional to the length of the resulting vector and is always positive (or negative if the diode is reversed). The control voltage U_c is only zero when U_1 and U_2 are equally great or if the sum voltage is below the threshold of the diode.

Whereas the synchronous demodulator supplies an output voltage at a certain phase difference, namely 90^0, whose amplitude is independent of the magnitudes of U_1 and U_{o1} (namely zero), the envelope demodulator, however, will cause U_c to react to any variation of U_1 and U_{o2}. The utilization of this circuit is, in many cases, limited by this as is the usable pull-in range of the local oscillator.



a) Circuit Diagram
U ... voltage
Uo2 = Auxiliary voltage from VFO
Uc = DC voltage resulting at the output

b) Vector Diagram at $\Delta\varphi$ 90° and $\Delta\varphi$ 120°
The DC voltage U_c results as the length of vector Uo2

c) Control voltage in dependence of the phase angle difference $\Delta\varphi$ between U1 and Uo2

Fig 3 Circuit Diagram Vector Diagram and Characteristic Curve of the Output Voltage U_c of an Envelope Demodulator

Frequency differences between U_1 and U_{o2} correspond to a continuous increase or decrease of the phase difference, which is also the case with the synchronous demodulator. This causes periodic fluctuations of U_c at the difference frequency

$$\left| f_1' - f_{o2} \right| = \left| f_1 - f_{o1} - f_{o2} \right| .$$

At greater deviations, the AC component of U_c will fall and the DC component will approximately reach the same value as when synchronized (phase difference $\Delta\varphi = 0$).

3.4. FILTERING OF THE CONTROL VOLTAGE

Since any variation of the control voltage U_c will cause a phase or frequency modulation of U_1, the control voltage must be carefully filtered. RF components from the control circuit oscillators, also the oscillators and intermediate frequencies of the receiver, can cause a modulation of the local oscillator frequency if they reach the varactor diodes. The same is valid for AF interference, noise, hum etc.

On the other hand, the filtering should not damp the circuit in such a manner, that the control voltage is not able to compensate for a rapid phase shift of U_1 (e.g. as a result of a load variation).

3.5. THE PHASE-LOCK RANGE

We have not described the actual synchronization process in the previous considerations. We would, therefore, now like to describe this process in more detail.

Firstly to the block diagram (Fig. 1). Let us assume that an increase of the control voltage U_c will cause the frequency of the local oscillator to also increase (the capacitance of the varactors is thus lower).

Statement: U_1 can only be phase-locked to $f_{o1} + f_{o2}$ if U_1 leads the voltage U_{o2} in the range $0°$ to $180°$.

Explanation: Let us assume that the circuit has been controlled in such a manner that U_1 leads the voltage U_{o2} by approximately $90°$. The voltage ratios of the discriminator circuit are given as a vector diagram in Fig. 3b (or Fig. 2b) , whereby the magnitude of the control voltage is given in Fig. 3c (or Fig. 2c). The local oscillator is now detuned so that its frequency slightly increases. The phase difference between U_1 and U_{o2} will then increase and U_c will decrease, as is shown in Fig. 3c (or Fig. 2c), e.g. for a phase angle of $120°$. However, the oscillator frequency is also reduced, and thus also the phase difference, until the original phase position is virtually achieved.

If the same consideration is made if U_1' lags U_{o2}, it will be seen that the phase difference will be controlled until U_1 leads U_{o2} by $0°$ to $180°$.

This merely means that the prerequisites for the correct operation of the circuit are only met if U_1' leads U_{o2}. Further information, such as the rejection of control oscillations, is not covered. Such considerations also

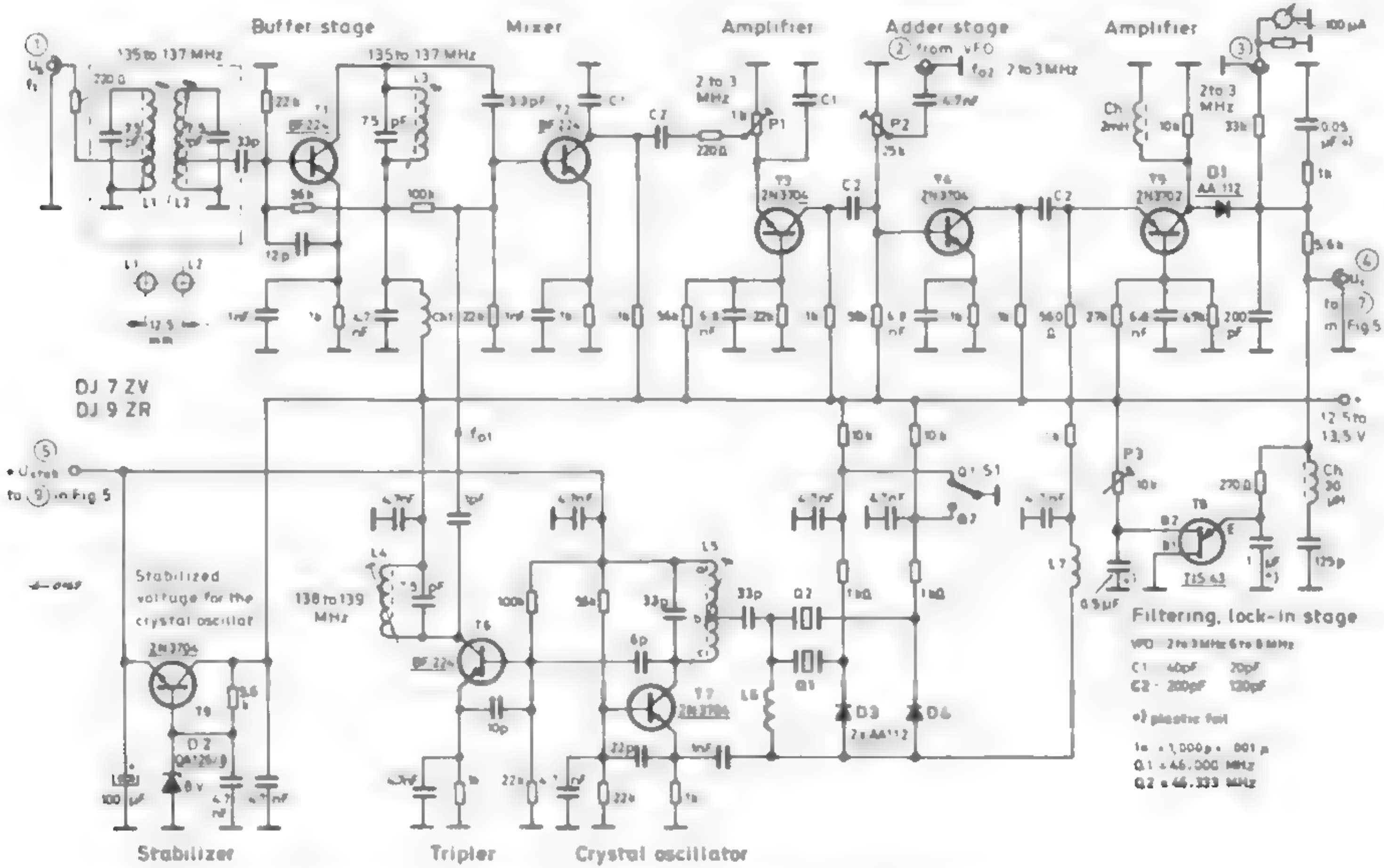


Fig 4 Circuit Diagram of the Phase-Locked Oscillator (Printed circuit board DJ 7 ZV 001)

show that the frequency f_1 can only jump to values higher than the nominal frequency if the circuit according to Fig. 3a is used and the control range of the varactor diode is not exceeded.

Of course, it is advantageous for the operating point and specifications of the varactor diode and the discriminator circuit to be selected so that a small phase variation causes a relatively great frequency deviation.

4. EXAMPLE OF PHASE-LOCKED OSCILLATOR CIRCUIT

Of the five different oscillator circuits developed, the circuit shown in Fig. 4 and 5 was found to be the most favourable. In this description, we will not discuss the cooperation between the individual stages and their importance in respect to the oscillator as a whole since the necessary information has been already given in Section 3 and is additionally shown in the block diagram (Fig. 1) associated with the circuit diagram. The pull-in range of the described phase-locked oscillator is greater than 2 MHz.

The frequencies given in the block and circuit diagram are only given as examples and can be modified if necessary. However, a complete transceiver using this phase-locked oscillator and the given frequencies will be described in a later edition of VHF COMMUNICATIONS.

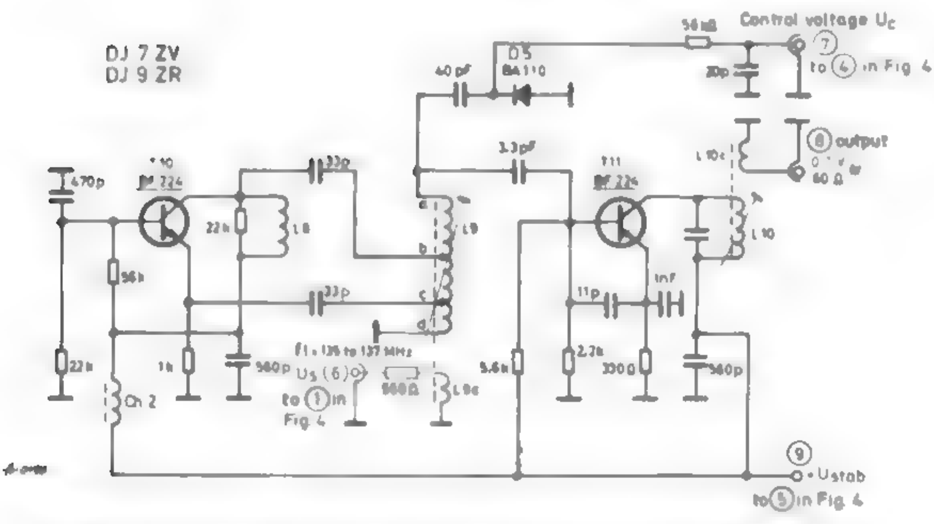


Fig 5 Circuit Diagram of the Local Oscillator for the Phase-locked Oscillator (f1)
(Printed circuit board DJ 7 ZV 002)

The local oscillator (Fig. 5) is equipped with the transistor T 10 and operates in a common-base three-point circuit. The varactor diode D 5 is connected to the resonant circuit via a 40 pF capacitor. The control voltage U_c is injected via a 56 kΩ resistor. If the DC voltage increases, the capacitance of the diode will decrease, thus causing the frequency to rise.

The oscillator is followed by an amplifier stage equipped with the low-reactive transistor T 11. A voltage of 700 mV - across 60 Ω - is available on the coupling link of the collector circuit for the transmit or receive mixers.

The input of the control circuit is connected to the local oscillator via a coupling link (L 9_c). This is followed by a two stage filter (L 1 and L 2) covering the frequency range of the local oscillator, and a common-emitter amplifier stage. The amplification of this stage is sufficient to drive the following mixer stage.

The crystal oscillator with transistor T 7 and the subsequent frequency multiplier stage with the transistor T 6 do not possess any special features. (Of course, it is also possible to equip the oscillator with a crystal oscillating at the first auxiliary frequency and to merely amplify the signal in the following amplifier stage). Since the tuning range of the variable frequency oscillator at the second auxiliary frequency is not sufficient to cover the whole 2 metre band, it is necessary to use two crystals which are then alternately switched into circuit via diodes D 3 and D 4. The operating voltage for the crystal oscillators is additionally stabilized with the aid of a pass-transistor (T 9) and a 8 V zener diode.

The RF voltage at the frequency f_{01} taken from the tripler stage is fed via a capacitor of 1 pF to the base of the first mixer with transistor T 2 and the signal from the buffer stage via 3.3 pF. A RC low-pass filter is to be found in the collector circuit of the mixer transistor which only allows the intermediate frequency f_1 (the transposed local oscillator frequency) to pass.

The trimmer potentiometer P 1 is to be found in the emitter circuit of the subsequent common-base stage. This trimmer is provided to adjust the IF current at the input of the following adding stage to a favourable value. The adding stage is designed to suppress any reaction of the second auxiliary oscillator voltage on the first mixer stage.

The tuning range of the second auxiliary oscillator (VFO for f_{02}) is 3.0 to 4.0 MHz in the case in question. The RF current fed from this oscillator to the base of the adding stage can be adjusted with the aid of P 2.

The total RF current fed to the base of the adding stage is the geometrical sum of the components obtained from the IF buffer stage and the auxiliary variable frequency oscillator. The corresponding vector diagram is shown in Fig. 6b. The resulting current is amplified in the subsequent transistors T 4 and T 5.

The phase comparison finally takes place in the rectifier circuit. The 200 pF capacitor represents a "storage capacitor" to which a loading resistor and a RC link are connected in parallel. This is followed by a series resistor and an absorption circuit for the intermediate frequency range as well as a RC link consisting of 270 Ω / 1 µF.

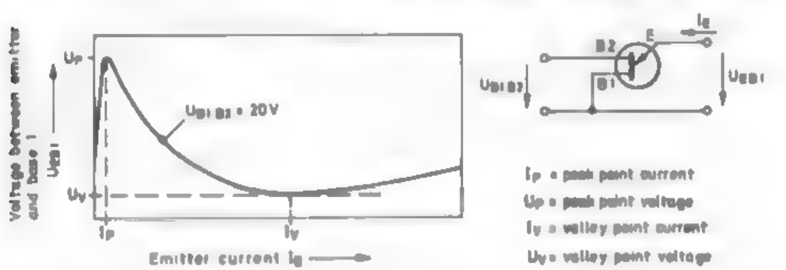
This filter network is calculated so that the IF voltage cannot be passed via the control line to the control diode. On the other hand, it ensures that the frequency control is made rapidly and that no oscillation of the control circuit can occur.

The lock-in circuit comprising the unijunction transistor T 8 is connected to the control line. The characteristic feature of the unijunction transistor is the fact that a change-over switching takes place between the emitter and base 1 when the emitter-base voltage exceeds a certain value which is dependent on the voltage between base 1 and base 2.

In the case in question, the two RF voltages simultaneously fed to the rectifier diode are adjusted in such a manner that the unijunction transistor switches when the frequency f_1 falls out of synchronization. The 1 μF capacitor will then discharge over the emitter-base 1 path, the control voltage will fall and the local oscillator will jump to a lower frequency. The capacitor is slowly recharged, the frequency f_1 and thus also f_1' will increase. As soon as f_1 has virtually the same frequency as f_{O2} and comes into the phase-locked range, the sum RF voltage and thus the rectified (and filtered) voltage will decrease and will balance itself. In the described circuit, the frequency of the local oscillator will be swept through the whole range in approximately 10 ms, whould the frequency fall out of synchronization.

We would now like to mention some characteristics of the unijunction transistor: Attention must be paid when connecting the unijunction transistor that the polarity of the two base connections is correct. Base 1 is connected to the negative and base 2 to the positive pole.

The characteristic curve is given in Fig. 6; the voltage between the emitter and the base is shown vertically and the emitter current horizontally.



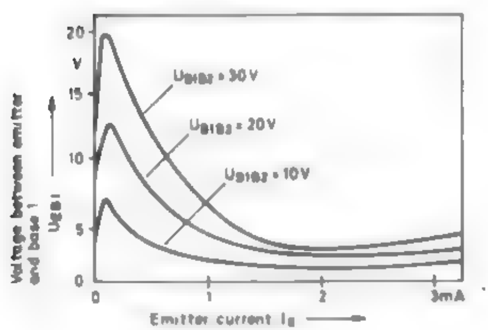


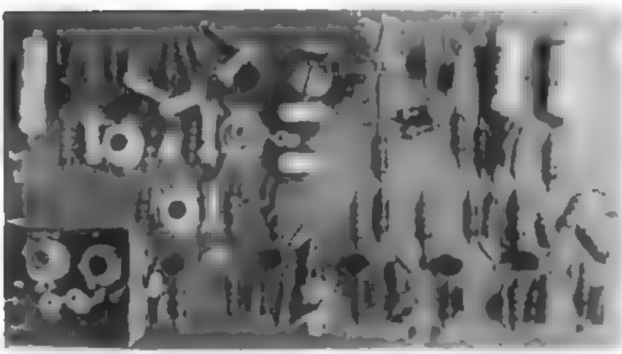
Fig. 6: Individual characteristic and Family of characteristics of a Unijunction Transistor

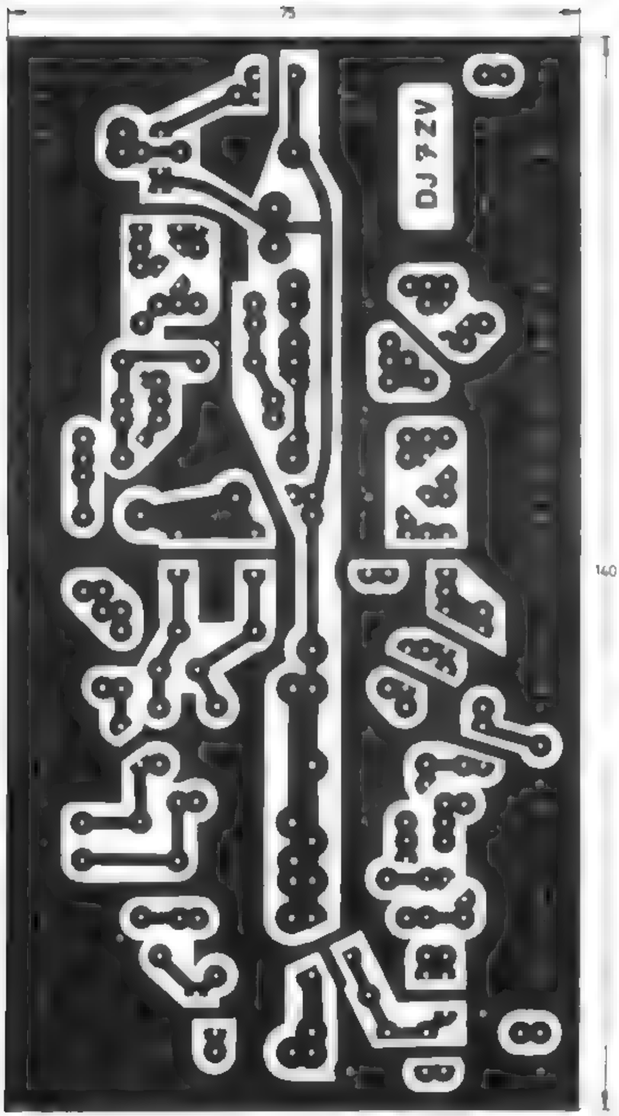
A small amount of emitter current will flow at an emitter - base 1 voltage of $U_{EB1} = 0$ which will attempt to positively charge the emitter; the current is in the order of a few microamps for the majority of unijunction transistor types. If a variable voltage source is placed between the emitter and the base and then increased, the current will firstly decrease to zero and will then increase, but will remain at a low level. When the supplied voltage finally reaches the peak of the curve ("switching voltage"), the negative portion of the characteristic will be reached: the emitter - base 1 resistance is so small that the voltage drops; a current results which is only limited by the impedance of the voltage source (if the component has not been destroyed previous to this). The return switching will not occur until the supplied voltage drops below the residual voltage level or until the current is reduced below the peak value. This peak value of the characteristic, which represents the switching point, can be varied by altering the voltage between base 2 and base 1.

In the case of the described circuit, the original condition of the unijunction transistor is obtained by discharging the capacitor. If the capacitance of this capacitor is greater than 0.2 µF, it will be advantageous to place a resistor into the emitter or base 1 lead.

3. MECHANICAL ASSEMBLY (Fig. 7 to 9)

The circuit according to Fig. 4 is accommodated on a printed circuit board as shown in Fig. 8 with the dimensions of 75 mm x 140 mm. The circuit of the free-running oscillator, as given in Fig. 5, is contained on the printed circuit board shown in Fig. 9. The latter is 50 x 85 mm in size. Figures 8 and 9 also include the component location plans as well as details regarding the components and values. The mounting of the components onto the two printed circuit boards should be made in conjunction with the two circuit diagrams and the location plans Fig. 8b and 9b. These plans contain all details with the exception of the coil data which is given in the following chapter. The photograph Fig. 7 shows the specimen board assembly built-up by the author DJ 7 ZV.





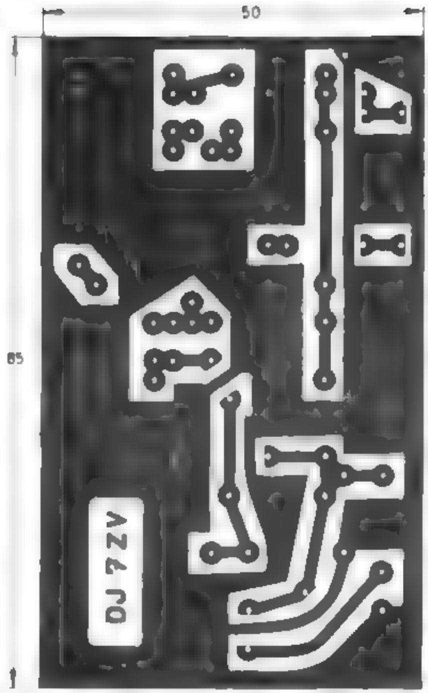


Fig. 9a

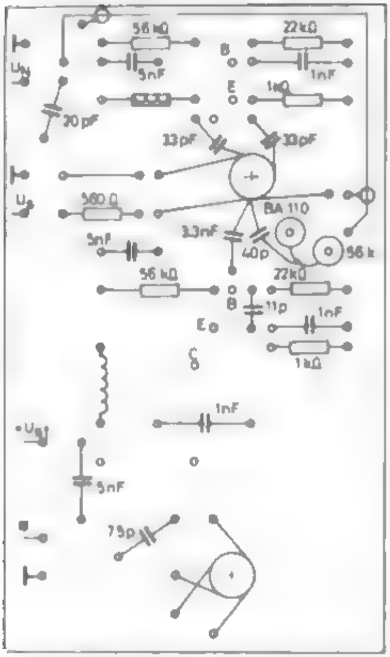
Printed Circuit Board
DJ 7 ZV 002
for the Local Oscillator
according to Fig. 5

Fig. 8a

Printed Circuit Board
DJ 7 ZV 001
for the Phase-Locked
Oscillator
according to Fig. 4

Coil Data

- L 1 = 6 turns of 0.8 mm dia. (20 AWG) PVC insulated copper wire wound on a 6 mm coil former equipped with a VHF core. Coil length approximately 8 mm. Coil tap 1 turn from the cold end.
- L 2 = 5 turns; wire, coil former and core as for L 1. Coil length approximately 6 mm. Coil tap 1 turn from the cold end. Spacing between the centre points of L 1 and L 2 = 12.5 mm.
- L 3 = 3.5 turns; wire, coil former and core as for L 1. Coil length approximately 4 mm.
- L 4 = L 3



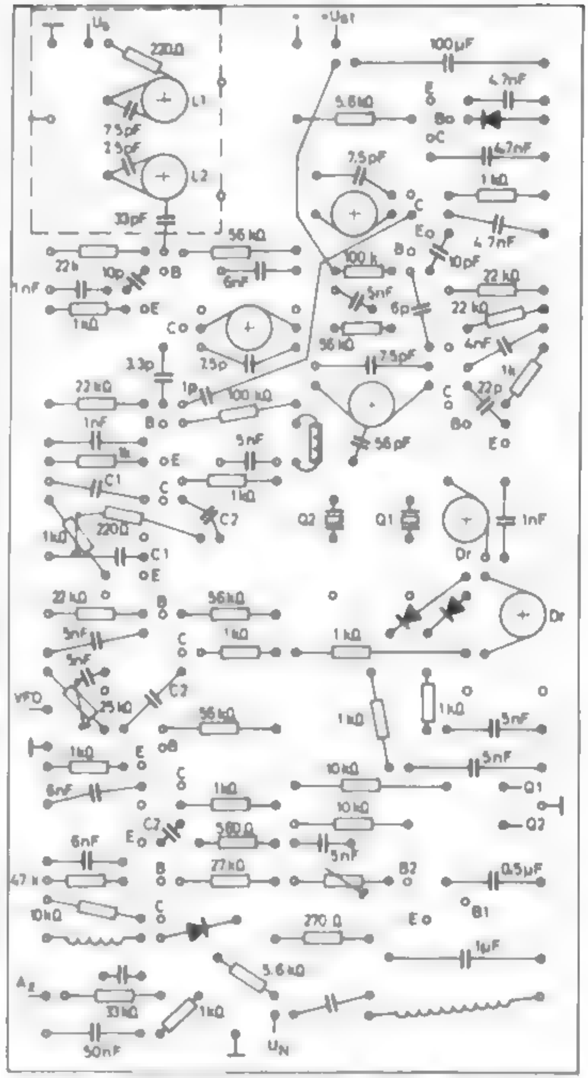


Fig. 9b

Component Location Plan of Printed Circuit Board DJ 7 ZV 002

Fig. 8b

Component Location Plan of Printed Circuit Board DJ 7 ZV 001.

L 5 = 7.5 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 6 mm coil former equipped with a VHF core. Coil length approximately 9 mm. Coil tap a - b = 0.5 turns.

L 6 = Choke, 15 turns of 0.3 mm dia. (29 AWG) enammeled copper wire wound on a plexiglass rod or other plastic former of 6 mm diameter.

L 7 = Choke, 46 turns of 0.3 mm dia. (29 AWG) enammeled copper wire wound on a 6 mm former similar to L 6.

L 8 = Choke, 23 turns of 0.3 mm dia. (29 AWG) enammeled copper wire wound on a 4 mm dia. resistor.

L 9	=	4 turns of 1 mm dia (18 AWG) silver-plated copper wire wound on a 6 mm coil former equipped with a VHF core. Coil length approximately 9 mm. Coil tap c - d = 1.7 turns.
L 9c	=	Coupling link, 1 turn of 0.8 mm dia (20 AWG) PVC insulated copper wire wound onto L 9.
L 10	=	3.5 turns; wire and coil former as for L 1. HF core.
L 10c	=	2.75 turns; wire as for L 1 wound between the turns of L 10.
Ch 1	=	15 turns of 0.3 mm dia (29 AWG) enammeled copper wire close-wound onto a 2.5 mm dia. resistor.
Ch 2	=	Wideband choke; ferrite bead, i.e. Ferroxcube Type VK 200 10/4 from Valvo or similar.

T 1	:	BF 224, BF 155 or 2N 3478
T 2	:	BF 224, BC 131, BC 109 or 2N 918
T 3	:	2N 3704, BC 131 or BC 109
T 4	:	2N 3704, BC 129 or BC 107
T 5	:	2N 3702 (PNP)
T 6	:	BF 224, BF 152, BF 173 or 2N 3478
T 7	:	2N 3704, BC 129 or BC 107
T 8	:	TIS 43 or 2N 489
T 9	:	2N 3704, BC 129 or BC 107
T 10	:	BF 224, BF 173 or 2N 918
T 11	:	BF 224 or 2N 3478
D 1	:	AA 112 or 1N 38
D 2	:	OA 126/8, 1N 430, 1N 1931 (Zener : 8 V, 250 mW)
D 3, D 4	:	AA 112, OA 182, 1N 118A or 1N 419
D 5	:	BA 110, BA 121 (Varactor: 10 pF at 2 V)

6. ALIGNMENT

The following measuring instruments are required for the alignment procedure, A dip meter, a means of monitoring the fixed frequencies f_{ol} and the output frequency as well as a DC multimeter, a suitably switched µA meter (high resistance meter with a 10 V FSD is sufficient) or a valve volt meter (VTVM).

Connect the VTVM to the control voltage connection 4 and short connection 5 to ground.

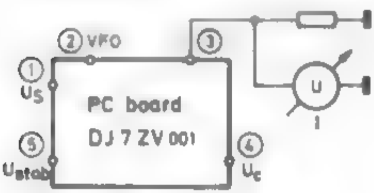
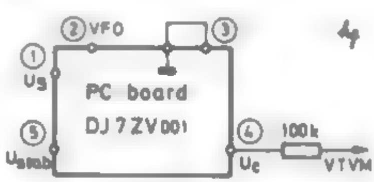


Fig. 10

Voltage Measurement using a voltmeter



Voltage Measurement using a VTVM

The alignment of the printed circuit boards is carried out in the following manner:

a) Equip the PC board DJ 7 ZV 001 (Fig. 8) with all components with the exception of the unijunction transistor and connect the operating voltage. A noise level should be observed at the control voltage connection (4).

b) Align the crystal oscillator and the frequency multiplier stage, after **which the frequency f_{o1} should be monitored.**

c) Assemble the printed circuit board DJ 7 ZV 002 according to Fig. 9. Connect a 60 Ω terminating resistor to the output and align the circuit so that the required frequency range f_1 of 135 MHz to 137 MHz is obtained at a control voltage of U_c = 3,2 V to 7,8 V. The control voltage U_c is taken from a battery during the alignment procedure. The frequency f_1 should now be monitored. This can be carried out by mixing f_1 with a 9 MHz signal to obtain the frequency of 144 MHz.

d) Build up the alignment circuit for the filters of the printed circuit board DJ 7 ZV 001 (Fig. 8). The VFO is still not connected. Tune the printed circuit (Fig. 9) from f_1 = 135 MHz to 137 MHz by altering the control voltage from U_c = 3,2 V to 7,8 V. Monitor the 144 MHz frequency resulting after mixing with the 9 MHz auxiliary frequency. The three stages located on printed circuit board DJ 7 ZV 001 comprising L 1, L 2 and L 3 are now aligned so that the required **bandpass characteristic is obtained.**

Meter I should indicate approximately 5 V. Switch crystals Q 1 and Q 2 of the f_{o1} oscillator to the correct position by operating switch **S 1.**

e) Disconnect the signal voltage connection (U_s) between the two printed circuit boards. Feed frequency f_{o2} from the variable frequency oscillator to connection (2) of printed circuit board DJ 7 ZV 001. The VFO should provide an RF voltage of at least 1.0 V over the whole frequency range. The reading on meter I should indicate 4.5 to 5.5 V **during this measurement.**

f) Connect the two circuit boards to another by means of a short coaxial cable. The unijunction transistor T 8 has still not be mounted onto printed circuit board DJ 7 ZV 001. Allow the frequency f_1 to be varied over the whole range from 135 MHz to 137 MHz by discharging the 1 µF capacitor at the emitter of the unijunction transistor T 8. The voltage reading on meter I should be in the range of 3,2 V to 7,8 V. If this is the case, the synchronization process is operating correctly.

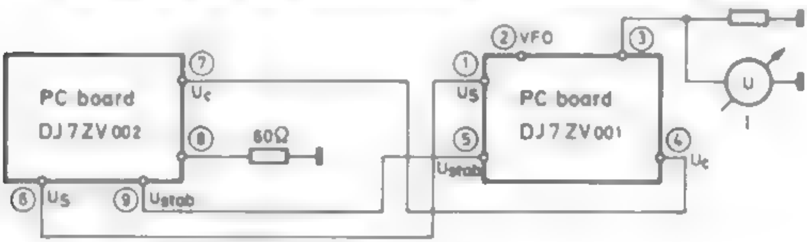


Fig. 11 Alignment Circuit (see Fig. 10)

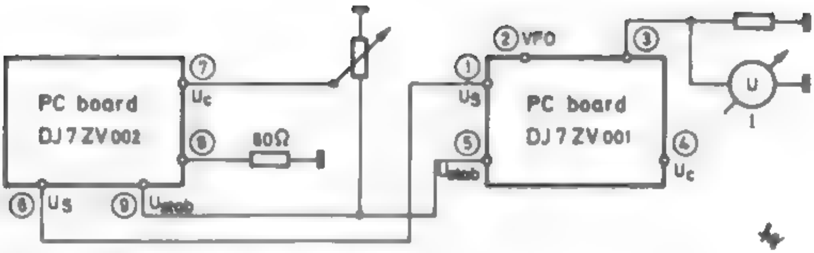


Fig. 12 Operational Circuit

A clean signal should now be heard in the receiver. The control voltage U_c is now varied by detuning the oscillator frequency f_1 (Fig. 5 or Fig. 9). The synchronized condition should, however, be maintained up to ± 1 V over the whole range of the VFO. It may be necessary to re-discharge the 1 µF capacitor at the emitter of T 8. It should now be possible for the output frequency f_1 to be varied over the whole of the 2 MHz wide frequency band by tuning the VFO (See Fig. 12).

g) The synchronization should be checked at all positions of the range; after which the signal voltage (U_s) connection is disconnected. The unijunction transistor T 8 should now be mounted on printed circuit board DJ 7 ZV 001. Adjust the 10 kΩ trimmer potentiometer in the base 2 circuit of T 8 so that the unijunction transistor correctly switches over the whole VFO range as a result of the voltage fed from the VFO to the demodulator (D 1).

Reconnect the signal voltage connection U_s. Adjust the voltage of the mixed signal so that the highest degree of synchronization is achieved (See Fig. 12). If the arrangement comprising circuit boards 001 and 002 does not synchronize; this will indicate that either one of the following conditions have not been observed or that the alignment given in Section 6 has not been carried out correctly.

Operating conditions for the phase-locked oscillator:
The crystal-controlled oscillator required for the generation of the fixed auxiliary frequency (frequencies) must oscillate to the value of frequency f_{o2} above the output frequency of the local oscillator. This is because the synchronization occurs from the lower frequency direction.

At an output frequency of f_1 = 135 MHz to 137 MHz and a VFO frequency of f_{o2} = 2 MHz to 3 MHz, the required fixed frequencies are f_{o1a} = 138 MHz and f_{o1b} = 139 MHz.

Further prerequisites for the correct operation of the phase-locked oscillator are:

a) The fixed auxiliary oscillator must generate a stable and clean signal.

b) All RF stages of the circuit must operate free of spurious signals.

c) The free-running oscillator for f_1 must only oscillate at the required frequency.

d) The variable frequency oscillator (VFO) for f_{o2} should be virtually free of harmonics and supply a constant signal voltage over the whole tuning range.

THE HB 9 CV ANTENNA FOR VHF AND UHF

by H. J. Franke, DK 1 PN

1. INTRODUCTION

This two element directional antenna was originally designed by R. Baumgartner, HB 9 CV for the HF shortwave bands (1) (2) and has been described - in a modified form for the two metre band - on several occasions in the magazine UKW-BERICHTE (3) (4). These descriptions were extremely popular because this relatively short antenna could be favourably used for fox-hunts, portable operation from mountain sites and for mobile operation. The fact that the antenna can also be dismantled is another advantage.

The following article describes firstly the results obtained with the two metre HB 9 CV antenna according to (3) and secondly two HB 9 CV antennas, one for the 70 cm band and the other for the 24 cm band.

2. THE HB 9 CV ANTENNA FOR THE TWO METRE BAND

The two metre antenna was assembled according to Fig. 1, which was also given in the description (3) and measured. In this description, the use of a trimmer capacitor between the connection point for the inner conductor and the inner conductor itself was recommended. On measuring the antenna impedance or standing wave ratio (SWR), it was observed that this trimmer is absolutely necessary in order to achieve a SWR of less than 3:1. In the most favourable position of the trimmer, a SWR value of 1.18:1 was achieved. This value was improved to 1.12:1 by slightly shifting the tapping point on the reflector. Since, however, only 4% of the power is reflected at a standing wave ratio of 1.5:1 (5), a value of 1.18:1 can be classed as extremely good for amateur considerations.

The most interesting of the measured values of the antenna is the gain. This is described as being approximately 5.5 dB in (2), 8 dB above a normal whip antenna in (3), to be 3 dB better than a three element mounted on a car in (4) (approx. 8 dB, ed.) and finally 6 to 7 dB better than a dipole in (6). Gain measurements which have been made by the author together with H. [illegible], DJ [illegible] VO and R. Lentz, DL [illegible] WR, have shown that a gain of approximately 4.5 dB over a half wave dipole was obtained with the two metre HB 9 CV antenna as well as for the later described 70 cm version. The side lobes at 90° and 270° were measured and found to be approximately 30 dB less than the value for the main lobe. The front-to-back ratio was, with approximately 14 dB, not extraordinary good.

3. THE HB 9 CV ANTENNA FOR 70 cm

During a 70 cm fox-hunt at the Schwabischer VHF Meeting in Ottobeuren, an HB 9 CV antenna with dimensions one third of the size of those given in Fig. 1 was used with success for direction finding. However, the antenna was measured afterwards and was found to have a SWR of 3:1 which

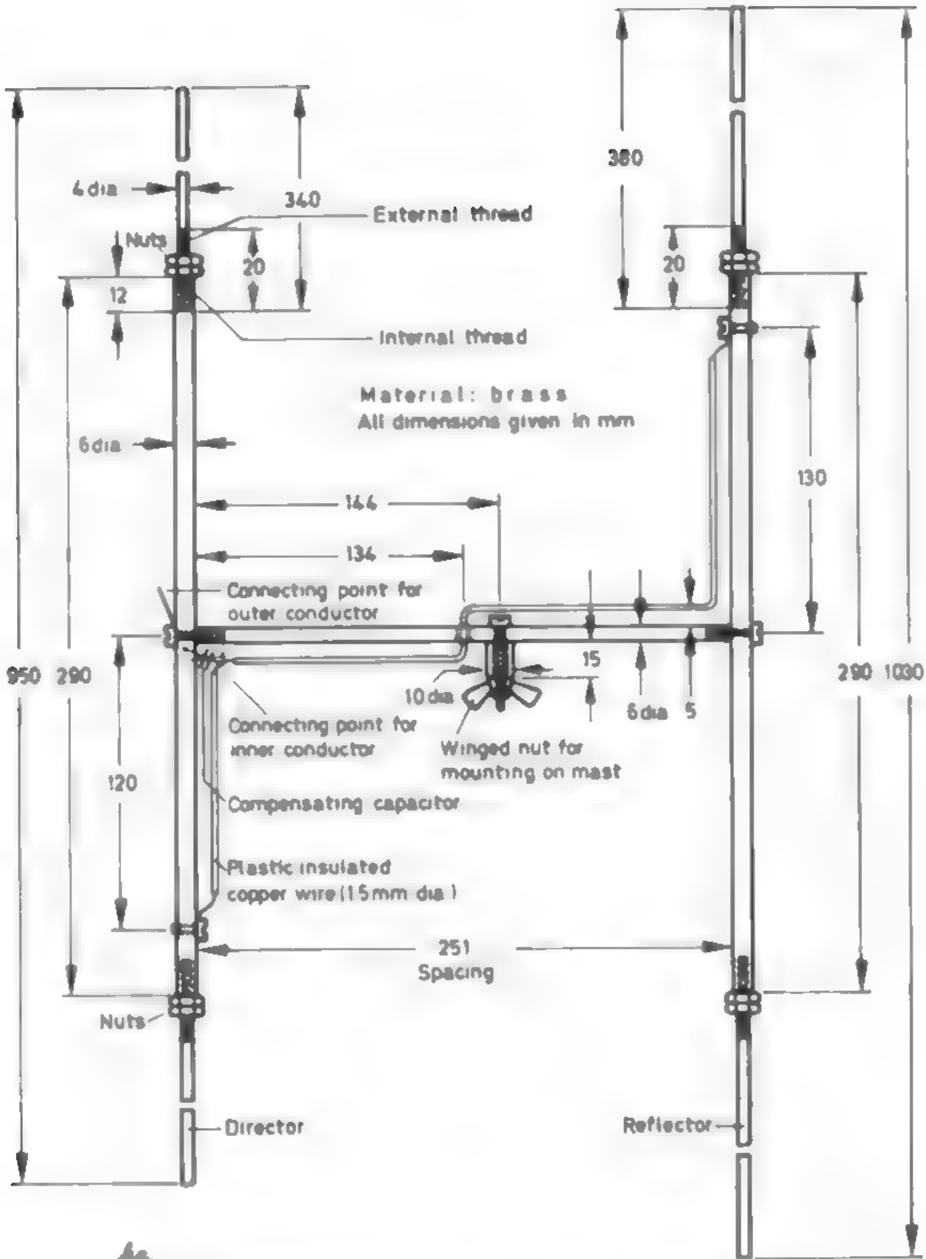
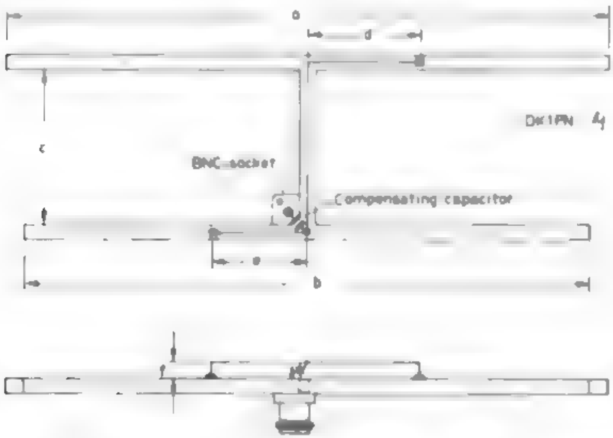


Fig.1 A Collapsible HB9CV Antenna for the 2 Metre Band

can no longer be classed as being tolerable. With the aid of a Z-g diagraph (an impedance measuring instrument manufactured by the company of Rohde u. Schwarz) other dimensions were found which enabled the same measuring values to be obtained as with the two metre antenna built-up according to Fig. 1. The feed line was, at the same time, formed in a different manner so that it could be more favourably matched and mounted to the antenna elements. By reducing the number of bends in the Z-shaped matching section - which is used to feed the two elements from the common feed line - the inductance was also altered which led to further deviations from the description (3). The final design and the dimensions of the 70 cm HB 9 CV antenna are given in Fig. 2. This small antenna cannot, however, be dismantled.



Dimens in mm	a	b	c	d	e	f	Compensat capacitor	Material Antenna	Match section
70 cm	338	306	83	57	53	6	≈ 3 pF	Brass tubing 5mm outer dia	Silver plated copper wire of 1mm dia
24 cm	112	103	27	19	17.7	2.6	≈ 1 pF	Brass 3mm outer dia	

1) The trimmer used by the author was the Valvo type 82001/12.5(0.65pF to 3.15pF). These trimmers are often used in the UHF tuners of TV receivers

Fig. 2 Form and dimensions of the HB 9CV antennas for the 70cm and 24cm band according to DK1PN

As already mentioned, the measuring values obtained with this antenna are practically identical to those of the two metre version. The SWR as a function of frequency was measured with the aid of a Z-g diagraph and is given in Fig. 3.

The compensating capacitor value of 3 pF only represents a nominal value. The attempt to replace the correctly adjusted trimmer with a fixed capacitor of the same value failed because only one of a number was found which enabled the low SWR to be maintained. After this capacitor was finally soldered into place with shorter leads, however, a far higher SWR was measured. The trimmer has been retained for this reason.

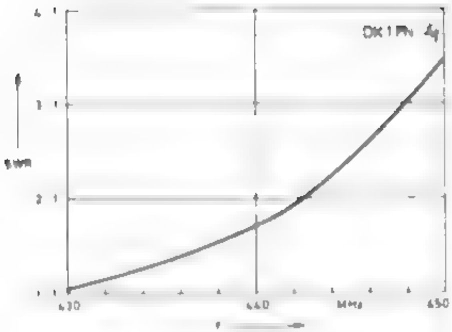


Fig 3 Standing wave ratio of the 70 cm HB 9 CV Antenna

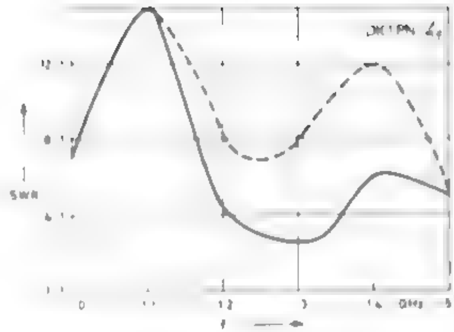


Fig 4 Standing wave ratio of the 24cm HB 9 CV Antenna
continuous line fixed capacitor of 0.7 pF
dashed line without compensating capacitor

The mechanical dimensions of this 70 cm antenna were also found to be extremely critical - and even more so for the 24 cm model described later. Changes of the material thickness can, for instance, only be carried out if a reflectometer (SWR-meter) is available which has been checked against an accurate terminating resistor.

4. THE HB 9 CV ANTENNA FOR 24 cm

The dimensions of the 70 cm model were divided again by three and the most favourable value of the compensating capacitor was determined. These values are given in the tables of Fig. 2. The SWR as a function of frequency was determined once again with the aid of the Z-g diagraph and is given in Fig. 4. The compensating capacitor has a value of approximately 0.7 to 1 pF. The same remarks are valid for this capacitor as for that of the 70 cm model. Since no favourable trimmer was available, the SWR measurement was carried out using a fixed capacitor with a nominal value of 1 pF. This could explain why the measured values (see Fig. 4) were not entirely satisfactory. Of course this antenna type, with a gain of approximately 4.5 dB at UHF, represents an antenna best suited for experiments or for short haul communications. The author would, however, firstly like to point out the success at 70 cm fox-hunts and to suggest the HB 9 CV antenna for use in stacked arrays.

5. THE MEASURING TECHNOLOGY

The antennas described in sections 3. and 4. represent the result of modifications, which have been carried out to obtain the most favourable standing wave ratio. Whereas the final measuring values were determined on a Z-g diagraph, a home-made reflectometer was used for the field experiments. The reflectometer was checked against a professional terminating resistor previous to these measurements, the reading was exactly 1 : 1 for both the 2 metre band and the 70 cm band.

A measuring arrangement as shown in Fig. 5 was used for the alignment process. The bandpass filter is used to suppress any harmonics or spurious signals from the (transistor) transmitter. This is necessary because any frequencies not corresponding to the pass band, will always be reflected from the antenna and would thus cause an increase in the SWR. This will be clearly seen if the SWR values fluctuate between different transmitters.

Fig.5 Arrangement used to measure the standing wave ratio DK1PN

The attenuator pad ensures that the signal component reflected from the antenna is not reflected again from the transmitter output. If this were the case, it would flow back and forth until converted into heat by the cable attenuation, which would also falsify the measured values. If the SWR approaches the value 1 : 1, the attenuator pad can be removed, however, the filter must be retained.

The measurement of the antenna gain was carried out after having aligned the antenna for the most favourable SWR. In addition to this, a professional attenuator was placed between the receiver and the antenna to be measured. This attenuator was adjusted for an insertion loss of 30 to 40 dB.

For your information, some of the values given in (5) are now given to show the reflected power in per cent as a function of the SWR.

SWR	1 : 1	1.2 : 1	1.5 : 1	2 : 1	3 : 1	5.87 : 1
Reflected power (%)	0	0.78	4	11	25	50

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144 MHz / 432 MHz TRANSVERTER FOR LOW POWER AND FIELD DAY APPLICATIONS

by L. Wagner, DL 9 JU

The idea to equip a 2 metre 70 cm varactor tripler with an additional crystal-controlled oscillator stage and, in this manner, to also use the varactor diode as the receive mixer, was incited by a lecture given by D. Vollhardt, DL 3 NQ, during the German VHF meeting held in Weinheim in 1965. This was followed by various discussions on the matter afterwards. The result was an extremely simple, but somewhat compromised, transverter which could be operated with any 2 metre transceiver with a transmit power of 100 to 500 mW, to form a 70 cm station. Higher powers could also be tripled by using a varactor with a higher power rating. However, the limited receive sensitivity (no preamplifier stage) would no longer represent a favourable relationship to the transmit power. The transverter is so small that it can normally be accommodated within the transceiver itself or, if this is not possible, it can easily be connected externally. The results obtained on operating from mountain sites in the Bavarian Mountain Field Day (BBT) have shown that 70 cm voice communication is possible over considerable distances using the described transverter (see Section 4).

1. CIRCUIT DIAGRAM OF THE 144 MHz / 432 MHz Transverter (Fig. 1)

The upper portion of the circuit diagram shown in Fig. 1 corresponds to a normal simple varactor tripler transposing a 2 metre signal to 70 cm. The components L 2 and C 3 form an idler circuit at a frequency of 288 MHz. To enable this tripler to be used as a converter in the receive mode, it is necessary to provide an oscillator to supply the diode with an auxiliary frequency of 288 MHz at a sufficient level. This is achieved with the auxiliary oscillator comprising transistor T 1. The oscillator operates in

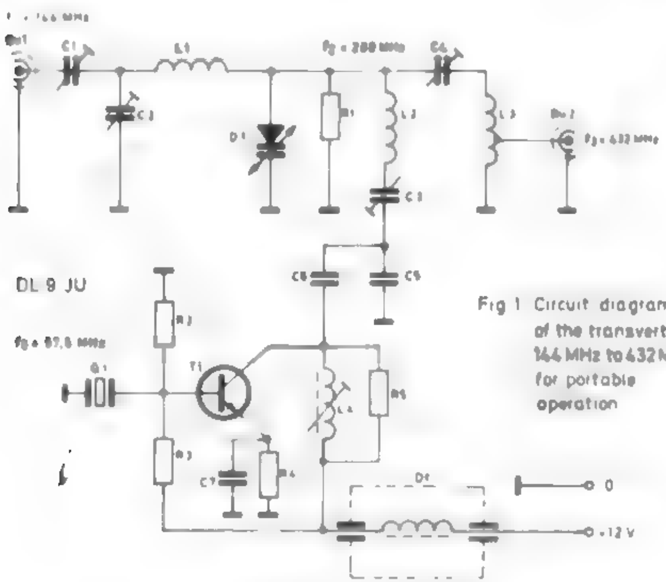


Fig 1 Circuit diagram of the transverter 144 MHz to 432 MHz for portable operation

a hybrid circuit (that is,none of the transistor electrodes are, in the RF sence, commonly used by two of the components connected) which operates at the frequency of the 57.6 MHz crystal Q 1. The 57 MHz resonant circuit, consisting of L 4 and C 6 / C 5, is to be found at the collector. The idler circuit with L 2 and C 3 / C 5 forms a filter circuit for the fifth harmonic (288 MHz) of the oscillator frequency which is connected to the oscillator via C 6. In this manner the power of the 5th harmonic is passed to the varactor.

The receive frequency of 432 MHz is mixed with the injected auxiliary frequency of 288 MHz to obtain the input frequency for the receiver of 144 MHz.

The attempt to quintuple the crystal frequency in the oscillator stage and to feed same to the auxiliary circuit - and in this manner to achieve a greater harmonic power - did not obtain the desired receive results. J. Reithofer, DL 6 MH, also came to the same conclusion. It is assumed that the circuit used produces a high harmonic content because the voltage dependent collector-base capacity allows the fundamental power to be converted into harmonic power, in a similar manner as the varactor, with low loss.

Special attention must be made in the selection of a suitable oscillator transistor T 1. The overlay transistor 2 N 3866 has shown itself to be well suitable for this purpose. Other transistors such as 2 N 708 and 2 N 2219 were found to be not suitable. Germanium UHF pre-amplifier transistors such as AF 139, cannot be used due to their low power dissipation. In order to obtain the highest harmonic content from the oscillator T 1, no additional capacitance may be connected in parallel to the collector stage inductance L 4 in addition to C 5 / C 6. The resistor R 5 serves to damp the collector stage. The 12 V power supply, which is only required to power the auxiliary oscillator T 1, is made via a feed-through filter or, at least, via a feed-through capacitor. Harmonics which could be injected into the stage from DC to DC converters are, if apparent, thus suppressed.

2. MECHANICAL ASSEMBLY OF THE TRANSVERTER

The mechanical assembly of the 144 MHz / 432 MHz transverter can be seen in Fig. 2, 3, 4 and 5. The circuit shown in Fig. 1 is built up on a non-etched epoxy board with the dimensions of 70 mm (2.75") by 100 mm (3.94"). The arrangement is not critical. The individual components are accommodated as shown in the fullscale drawing Fig. 2b.

After having made the cutouts for the two coaxial sockets, the transistor socket, the crystal socket, the feed-through filter, the coil formers as well as the mounting holes for the four trimmers and the four mounting bushings (see Fig. 3), the assembly may be commenced. All ground connections should be made to the coated side of the circuit board. The mounting of the varactor diode should be made with the shortest leads, remembering to cool the connection leads with a pair of pliers or similar during the soldering procedure. Four spacing bushings - see Fig. 3b - are to be found on the coated side of the circuit board in addition to the components. The complete arrangement can be covered with an aluminum plate as shown in Fig. 3a which contains all the cutouts necessary for the alignment procedure.

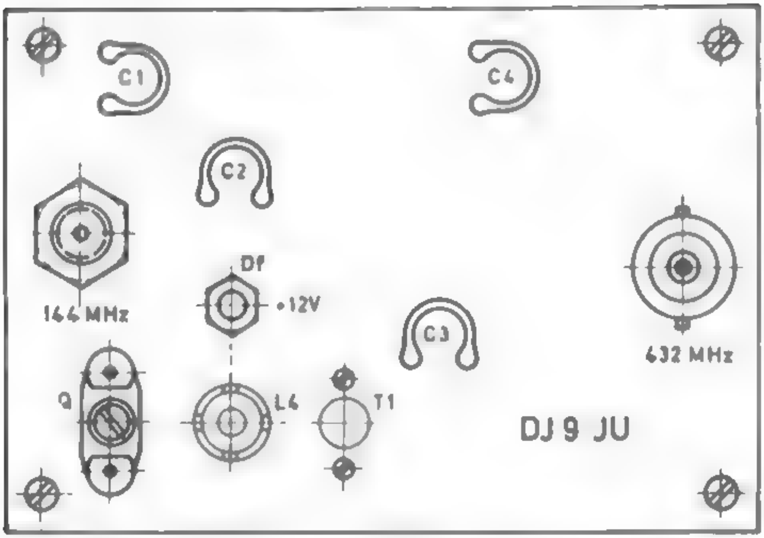


Fig. 2a: Transverter 144/432 MHz from above

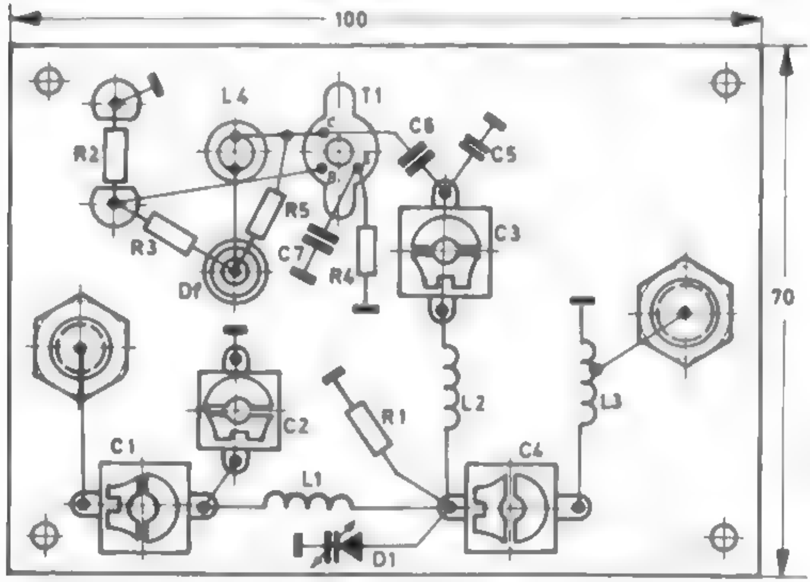


Fig. 2b: Mechanical assembly of the tranverter
from below

2.1. COIL DATA AND COMPONENTS

L 1 5 turns of silver-plated copper wire, 1.2 mm (17 AWG) dia. wound on a 10 mm former, self-supporting, approx. 15 mm long.

L 2 4 turns of silver-plated copper wire, 1.2 mm (17 AWG) dia. wound on a 6 mm former, self-supporting, approx. 8 mm long.

L 3 3 turns of silver-plated copper wire, 1.5 mm (15 AWG) dia. wound on a 6 mm former, self-supporting, approx. 8 mm long and with tapping 0.75 of a turn from the cold end (ground).

L 4 7 turns of cotton covered enamelled copper wire, 0.3 mm (29 AWG) dia. on a coil former of 6 mm dia. (see Fig. 6), HF core.

Q 1 57,6 MHz crystal in a ceramic holder.
C 1, C 2, C 3 Air spaced trimmers 11 pF max.
C 4 Air spaced trimmer 5 pF max.
C 5 Ceramic disc capacitor 100 pF
C 6 Ceramic disc capacitor 10 pF
C 7 Ceramic disc capacitor 15 pF

Skt. 1, Skt. 2 BNC coaxial socket UG-1094/U, Z = 52 Ω

D 1 Varactor diode BA 121 or BA 149 AEG-Telefunken or Intermetall BA 110. (BA 121: C_j = 10 pF at U_R = 2 V PIV = 30 V Q = 600 U_R = 2 V and f = 30 MHz max. diss. = 250 mW case = DO 7).

T 1 2 N 3866 from RCA
R 1 30 kΩ; R 2 2.7 kΩ; R 3 10 kΩ
R 4 270 Ω; R 5 3.3 kΩ

All resistors are rated at 0.1 watt.

3. ALIGNMENT OF THE 144 MHz / 432 MHz TRANSVERTER

The alignment of the transverter is made in the following manner: Firstly diode D 1 is low-inductively bridged with a metal strip. The idler circuit consisting of L 2 and C 3, C 5 is aligned, using a signal generator or a dip meter, to the frequency of 288 MHz, after which the bridge is removed. The trimmer C 3 should, after this, not be altered.

The 2 metre transceiver is connected to the 144 MHz input socket Skt. 1, and socket Skt. 2 is terminated with a 52 Ω resistor as well as being connected to a 432 MHz indicating circuit (e.g. an absorbtion circuit with diode and meter). The transmitter should now be switched on and capacitors C 1, C 2 and C 4 aligned for a maximum reading on the indicator. This procedure should be repeated several times due to the interaction between the individual circuits.

After this the transceiver is switched to receive. A 70 cm signal is now fed to socket Skt. 2, whereby the harmonics of a 2 metre transmitter are usually sufficient. The inductance L 4 should now be aligned for maximum receive volume. A favourable compromise between the most optimum tripling in the transmit mode and the best reception in the receive mode should now be determined by capacitors C 1 and C 2. The alignment is with this completed.

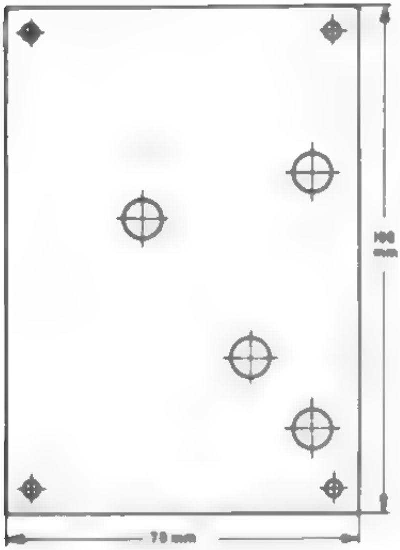
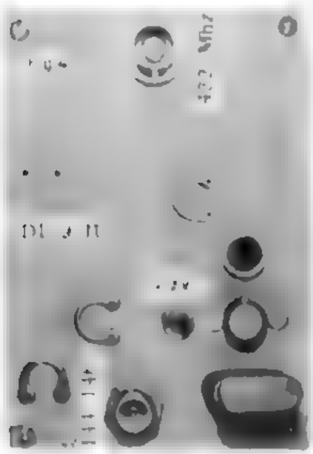
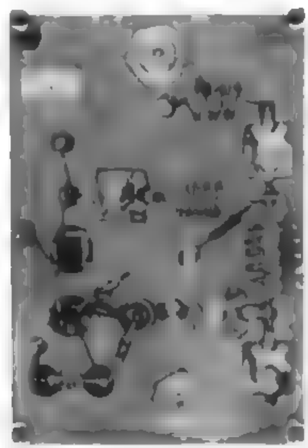


Fig. 3b
Spacing pieces, aluminium, 4 required

Fig. 3a
Aluminium plate for the lower cover 1.5 thick. Dimensions 100 x 70 mm (3.94" x 2.75")





4. OPERATIONAL RESULTS AND NOTES

The noise figure of the 144 MHz / 432 MHz transverter was measured in conjunction with a 2 metre transistor converter with a noise figure of approximately 10 dB. The efficiency of the tripler in the transmit mode was in between 40 and 50%. The described transverter was used by the author for 70 cm operation from mountain sites in the Bavarian Forest, whereby voice communication was made over distances exceeding 150 km (95 miles). The output power of the 2 metre portable station was approximately 200 mW which means that the maximum possible power on 70 cm was not more than 100 mW.

A SOLID-STATE CONVERTER FOR 24 cm

by R. Lentz, DL 3 WR

The following article has been written more extensively in order to discuss the problems encountered in conjunction with diode mixer and varactor diode multiplier stages. We hope that the article will be informative and also of interest to those readers who do not intend to build the converter or only wish to copy some particular portion of the design.

1. INTRODUCTION

The following 24 cm converter is exclusively equipped with silicon semiconductors and can thus be operated from a 12 V source. Since the total current drain is only 20 mA, the converter can be run from the batteries of a transistorized portable transceiver. This and the fact that the unit only weighs 380 g (13. oz.) makes the converter extremely attractive for field day and operation from mountain sites, where no AC line facilities are available or desired.

The mechanical assembly is extremely simple and requires no precision metal work. The carefully cut chassis plates can be soldered together on the work bench, after which it is merely necessary to drill a few holes and to commence assembly.

In spite of its simplicity, the transistorized 24 cm converter is technically advanced, which is indicated by the measured noise figure of 9.5 dB.

The frequency range of 1296 to 1298 MHz is converted to 28 to 30 MHz with the aid of the 84.533 MHz crystal. Slight variations from this frequency range are possible by use of a different crystal frequency. A conversion to a first IF of 144 to 146 MHz is, however, not recommended because the low noise figure will not be achieved. Due to the design of the described converter - no RF stage, mixer diode with conversion loss and a sensitive preamplifier - the absolute sensitivity of the converter is essentially determined by the absolute sensitivity of the IF preamplifier. This can be calculated as follows:

$$\boxed{NF_{tot} = NF_{mix} + \frac{NF_{IF} - 1}{PG_{mix}}}$$

Whereby: NF_{tot} is the total noise figure

NF_{mix} is the noise figure of the mixer diode

NF_{IF} is the noise figure of the IF preamplifier

PG_{mix} is the power amplification of the mixer stage. i.e. a conversionloss of 6 dB. means $PG_{mix} = 0.25$

Thus a lower absolute sensitivity of the IF preamplifier obtained by using a higher first IF has the effect of increasing the total noise figure by multiplying itself with the conversion loss of the mixer diode.

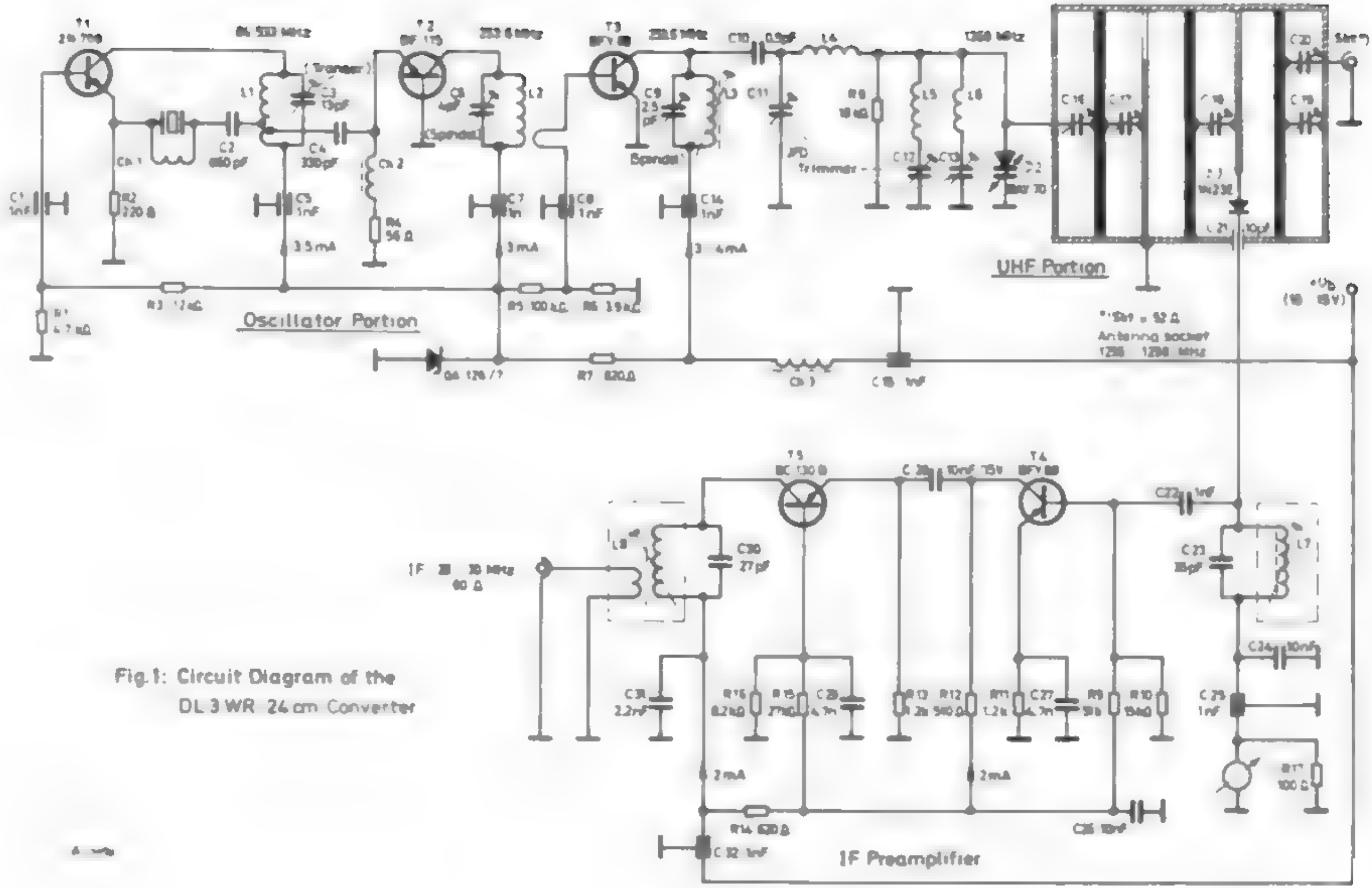


Fig.1: Circuit Diagram of the DL 3 WR 24 cm Converter

The UHF portion of the converter, i.e. the three resonant line circuits and the mixer diode, is based on a circuit given in QST (1). A similar design was also given in an article in UKW BERICHTE (2).

This UHF portion was modified by C. Sondhauss, DJ 2 LJ for capacitive coupling, the provision of a varactor quintupler and otherwise equipped with tubes. Due to the low noise figure obtained with tubes in the 30 MHz range, the absolute sensitivity was so great as to allow a total noise figure of 7.7 dB.

Unfortunately the DJ 2 LJ converter was never published but the UHF portion and the varactor quintupler have been taken over by the author in the described converter.

We would also like to mention the 24 cm converter published in the Telefunken Taschenbuch (3) which possesses two grounded grid amplifier stages and a tubed mixer with the Telefunken tube 8255, an improved version of the EC 88. The sensitivity of the IF amplifier (HF receiver) has practically no effect on the total noise figure with this rather extensive circuit and a first IF of, for instance, 145 MHz could be used. The absolute sensitivity of the converter is 9.5 dB.

2. CIRCUIT DESCRIPTION (Fig. 1)

2.1. GENERAL DETAILS

The converter consists of three main sections: the oscillator portion, the UHF portion and the IF preamplifier. The oscillator portion commences with an overtone crystal oscillator at a frequency of 84.533 MHz which is followed by a tripler stage, multiplying the oscillator signal to 253.6 MHz and a subsequent buffer amplifier stage. The amplified signal at a level of more than 10 mW is then fed to a varactor quintupler, whose output signal of 1268 MHz is decoupled via a resonant line transformer and passed to the mixer diode. The overtone crystal frequency is thus multiplied a total of 15 times.

The input frequency of 1296 to 1298 MHz is capacitively coupled from the input connector to the input resonant line and is also inductively fed to the mixer diode.

The intermediate frequency generated in the mixer stage is passed to a cascode preamplifier circuit. As is well known, it is necessary for the cascode circuit to have the first stage in a common emitter and the second stage in a common base configuration. No use has been made of selective links between the two stages (RC link).

2.2. CIRCUIT DETAILS

2.2.1. OSCILLATOR PORTION

The oscillator operates in a common base circuit and the crystal holder capacity is neutralized by choke Ch 1. The coil tap on L 1 for the feedback and coupling out of the signal is adjusted for the maximum drive to the tripler transistor (maximum collector current). L 1 has not been

provided with an alignment core in order to increase the frequency stability and the appropriate trimmer C 3 is an air-spaced type. Transistors 2 N 708 (BSY 19), 2 N 914 (BSY 21), 2 N 918 or even the LF types such as BC 108, BC 130 etc. can be used in this stage.

A transistor BF 115 in a common-base configuration is used in the tripler stage and operates in class C. A resistor in the emitter circuit compensates for slight variations of the operating point and drive conditions. This resistor also adjusts the angle of current flow to the most favourable value for the drive level (highest harmonic content).

The oscillator and tripler are operated with a stabilized voltage, obtained by use of 7 volt zener diode. This is extremely necessary for reasons of frequency stability and also ensures a constant drive to the subsequent buffer amplifier stage.

Only a transistor with a high transit frequency and a low retroaction can be used for the 253 MHz amplifier stage and a BFY 88, BFY 90 or 2N 3478 with a transit frequency of 900 MHz and a retroaction of 0.23 pF (due to integrated insular screening) was chosen by the author. This transistor may only be used in a common emitter configuration. The transistor is adjusted for class B operation via a voltage divider fed from the stabilized operation voltage so that the collector current just commences to flow under no-signal conditions. This adjustment results in the highest output power at low input levels. When matched to 60 Ω, this stage will provide an output power of 12 mW at 12 V operating voltage which was sufficient to obtain a diode current of up to 2 mA across the varactor quintupler of the described converter.

The quintupler is high impedance fed via capacitor C 10 and the matching is made with inductance L 4 and trimmer capacitor C 11. The idler circuit comprising L 5 and C 12 is aligned to the 2nd harmonic 507.2 MHz and the idler circuit with L 6 and C 13 to the 3rd harmonic 760.8 MHz. Tubular glass trimmers were used in the prototype converter which had the disadvantage of being somewhat expensive but other trimmer capacitors with the corresponding capacity values should be equally suitable.

A commercial varactor diode BAY 79 manufactured by Telefunken, which was contained in a special holder, was used by the author. The junction capacitance is 5 pF at a reverse voltage of 2 V and the breakdown voltage is 35 V. These specifications are also met with the currently available Telefunken diode BAY 70, which should be connected using short leads between the junction of components L 4, L 5, L 6, R 8 and C 16 and ground (see Fig. 5).

2.2.2. UHF PORTION (Fig. 4)

The three coaxial circuits of the UHF portion are slightly shortened half-wave (λ/2) circuits. The voltage lobe is thus in the centre of the lines where they can be capacitively shortened and tuned. The oscillator frequency is fed at a somewhat lower impedance to a point approximately half-way towards the cold-end where it is also capacitively coupled. The oscillator frequency transformer is coupled via the cut-out between the two reasonant lines (see Fig. 2 and 5). The mixer diode is also to be found in a similar cut-out in the partition between the input circuit and the secondary of the transformer. The thick end of the diode protrudes into the IF preamplifier compartment and is blocked for UHF frequencies by a home

made 10 pF plate capacitor (see Fig. 4). This excellent conception of the author of publication (1) has solved the critical coupling and mounting problems encountered with mixing diodes. The UHF portion was so easy to tune that no modifications have been made on the basic design.

Measurements carried out by C. Sondhaus, DJ 2 LJ have shown that the diode series 1 N 23. offered no lower noise figure at 1.3 GHz than 1 N 21 . types. The author uses a 1 N 23 E diode which is operated with a current of 0.8 to 1 2 mA. This current is then indicated on a meter shunted for 2 mA f.s.d.

2.2.3. IF PREAMPLIFIER

The input of the IF preamplifier must be so dimensioned to allow both the mixer diode its optimum IF impedance of 300 to 500 Ω and the most optimum source impedance of the input transistor, whereby the input impedance of the transistor acts as the terminating impedance for the diode and the impedance of the (current carrying) diode as the source impedance for the transistor. Since these impedance values lay in the same order of magnitude, a transformation process was not used. It could, however, be possible that a more exact matching at this point would improve the absolute sensitivity.

The input circuit should possess a highest possible unloaded Q since all equivalent resistances at this point have an adverse effect on the noise figure. In extreme cases it may be adviseable to form inductance L 7 in a different manner and to additionally isolate the base voltage divider from the critical high point of the first circuit with a RF choke.

The L/C ratio of the first IF stage has, in conjunction with the UHF blocking capacitor C 21 and the transistor input capacitance, been so chosen that the circuit is damped to a Q of 3 by the output impedance of the diode and the input impedance of the transistor. This corresponds to a bandwidth of approximately 10 MHz. The alignment is thus uncritical and it is only necessary to alter the tapping on inductance L 8 to modify the converter to receive wideband signals such as FM or amateur television transmissions.

In the case in question, the output circuit was aligned to a bandwidth of approximately 3 MHz, which is valid for a termination of 60 Ω. A greater bandwidth can be obtained by using a lower transformation ratio (more turns on the secondary).

The transistor BFY 88 , BFY 90 or 2 N 3478 is also used in the input stage of the IF preamplifier where the major portion of the sensitivity is determined (see Section 2.2.1.). Transistors having a lower transit frequency such as BF 173, BF 224 and BF 115 would also be suitable but with a slightly lower absolute sensitivity.

Since silicon planar transistors such as BC 130 or BC 108 possess transit frequencies of approximately 250 MHz, these types can be used in the less critical second stage of the cascode circuit. The large retroaction capacity does not have any adverse effect in this circuit.

3. MECHANICAL ASSEMBLY (Fig. 2 to 10)

3.1. MAIN CHASSIS INCLUDING UHF PORTION AND VARACTOR MULTIPLIER

The converter is built-up on a glass fibre reinforced 1.5 mm epoxy board with a copper coating on both surfaces. This material can be easily processed at home and can be simply soldered with the aid of a small soldering iron. The exception to this assembly is the UHF portion whose wall pieces and partitions are made from brass plate and the inner conductors from brass tubing. This is to ensure that RF currents at the ends of the inner conductors and around the cut-outs in the partitions 1, [illegible] and [illegible] can flow freely. The brass plate is 0.5 mm thick.

Since it is impossible to use bent chassis portions when assembling on printed circuit boards, the already tailored metal pieces are soldered directly on to the non-drilled main board, which later forms the upper side of the converter. The walls are provided with slots to enable the assembly to be simplified and to give more support before soldering. Fig. 2 shows the assembly of the main chassis, Fig. 3 the mechanical dimensions of the partitions and Fig. 4 the UHF portion of the converter with the corresponding mounting instructions. Table 1 contains all details regarding the casing pieces and mechanical accessories.

The assembly is commenced with the innermost partition 2 so that the maximum space is available for further soldering. After soldering the side, end and partition pieces to the main chassis, the UHF input socket Skt 1 is soldered on the inside after having drilled the corresponding hole. It is also possible to screw the socket but care must be taken that a good ground contact is made inside the coaxial circuit. When using printed circuits coated on both surfaces it is important that the ground points are made on the correct side.

The next item to be mounted is the coupling loop C 20 which is soldered to the inner conductor of the UHF input socket Skt 1 (see Fig. 4B). The three inner conductors, which have been drilled through with a [illegible], are now placed into end piece 2 and partition 4 to find the hole positions for the tuning trimmers which are located directly under the centre holes of the inner conductors. The three holes can now be drilled in the main board. It is now possible to mount the trimmer capacitors C 17, C 18 and C 19 which are then soldered on the inside surface. The spindles of the trimmer capacitors are now removed, placed into the centre holes of the inner conductor and the spindles once again screwed [illegible] through the centre hole of the inner conductors. The inner conductors are now adjusted so that the spindles are not touched and then soldered.

Following this, it is necessary to drill a hole in the main board for the ceramic soldering tag, on to which the coupling flap C 16 is mounted (see Fig. 4 C). The solder tag receives the flap instead of the original tag and is installed into the converter. The holes are now drilled in the main board for the three trimmers of the varactor quadrupler (see Fig. [illegible]), for the four feed-through capacitors and for the low capacity feed-through (see Fig. 8). The cut-out for the meter should also be prepared (see Fig. 5). The contact spring of the mixer diode is now soldered into the cut-out in the first partition and the plate for capacitor C 21 prepared for mounting. It is now possible for the converter chassis to be silver plated. In the prototype converter the cover and the sub-chassis were not silver plated.

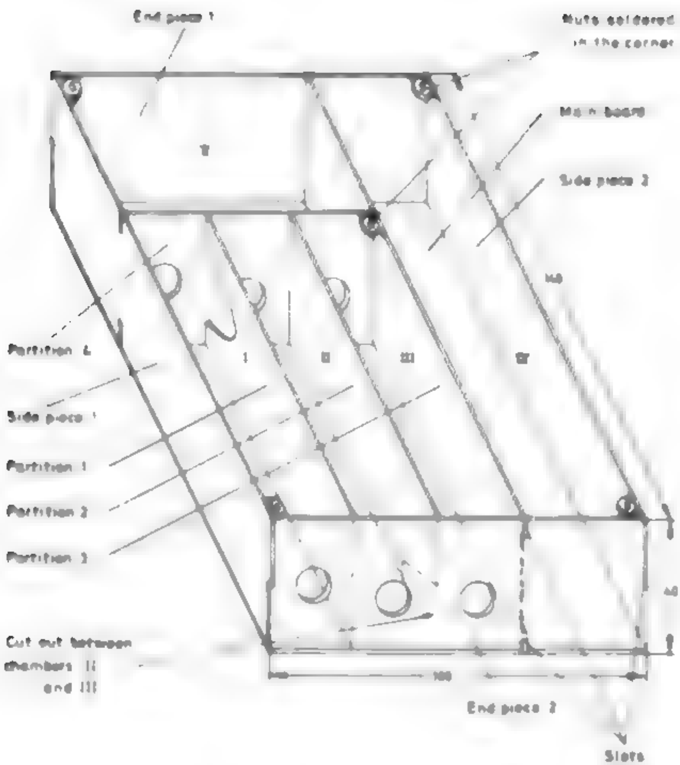


Fig. 2 Main Chassis of the 24 cm Converter showing chamber I to V (DL 3 WR)

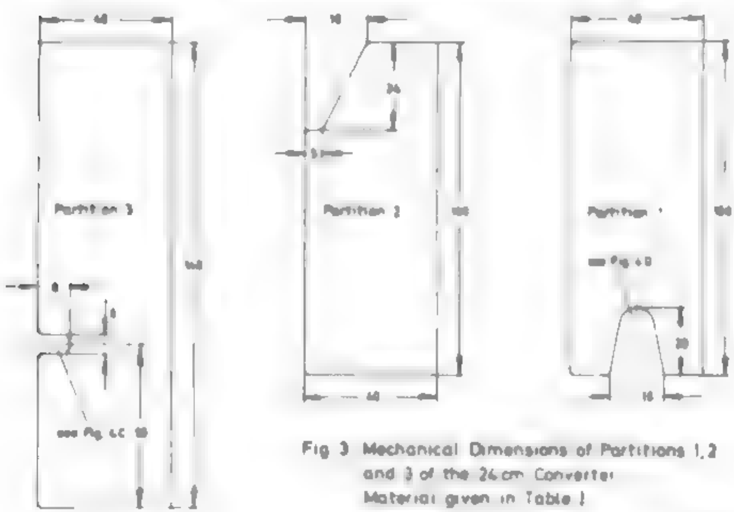


Fig. 3 Mechanical Dimensions of Partitions 1, 2 and 3 of the 24 cm Converter
Material given in Table 1

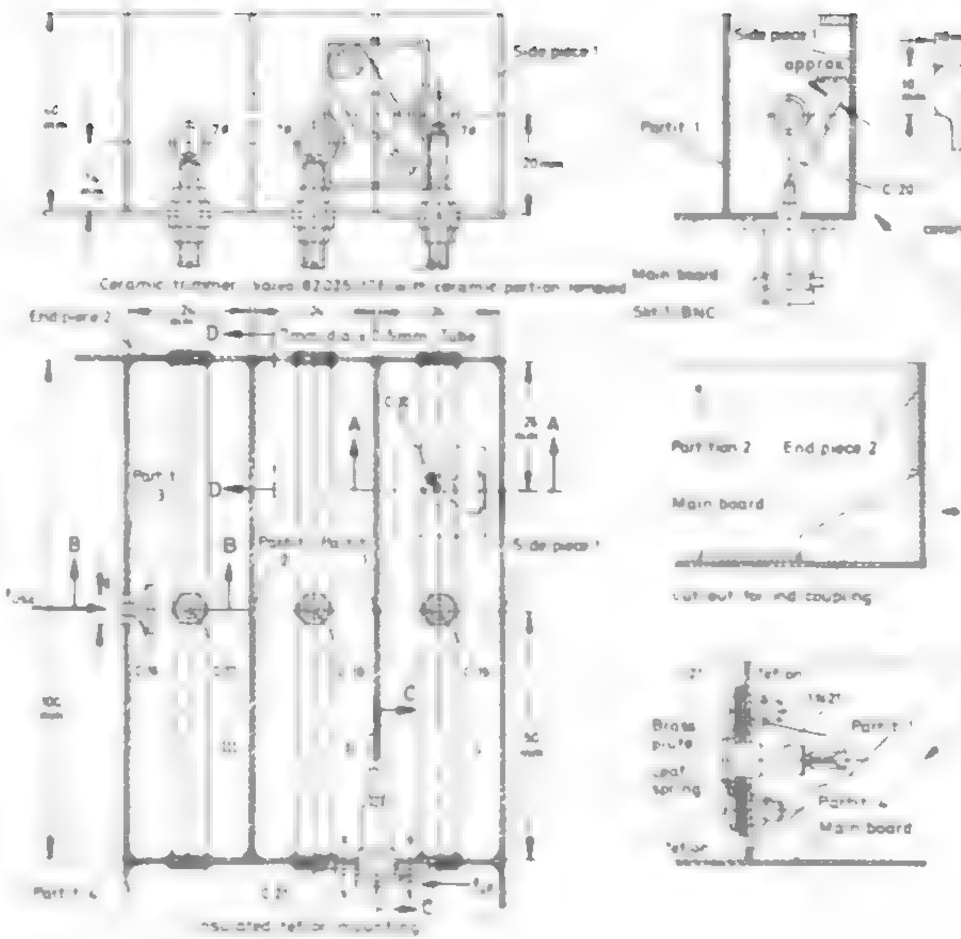


Fig 4c Sectional View B B
C16 and C17 in UHF
Portion

Fig 4b Sectional View A-A
Skt 1 and C20 in UHF-
Portion

Fig 4e Sectional View D D
Cut out in Partit 2

Fig 4d Sectional View C-C
Mounting of the
Mixer Diode

Fig 4a (left hand drawing)
UHF Portion of the
DL 3 WR 1296 MHz
Converter
(Chambers I, II and III)

The solderable feed-through capacitors are now soldered to the inside of the converter and the IF feed-through pressed into place.

The varactor multiplier is mounted in chamber IV beside the coupling flap, in order to obtain the shortest possible connections to the resonant line transformer. The star point of the circuit is the hot end of the diode, there being no connecting leads. The three inductances are arranged in different directions in order to obtain a decoupling effect. The resistor R 8 should be as low as possible and its value can be varied to establish the most optimum value. See Fig. 5 for further details.

The oscillator portion, excepting quadrupler, and the IF preamplifier represent two small sub-assemblies and are separately assembled, prealigned and thereafter mounted on to the main chassis.

3.2. OSCILLATOR PORTION (Fig. 6)

This separately assembled sub-assembly comprises the crystal oscillator, its tripler and the 2[illegible] MHz tripler amplifier stage. It is built up on a small piece of printed circuit board, which is copper coated on both surfaces (see Fig. [illegible]). A partition made from the same material, on which transistor T [illegible] is mounted, separates the input and output of the 2[illegible] MHz amplifier stage. Of course, brass plate could also be used for this sub-assembly, and with enough experience, be modified into a printed circuit board.

The order of assembly for this sub-assembly is briefly described as follows: Saw the plate and screening pieces to size, drill all the necessary holes, solder in the partition and then solder the feed-through capacitors into place (the upper side of the printed circuit board is used as ground for all RF ground points). Solder two screws (approx. 15 mm long) to suitable positions on the printed board and drill the corresponding holes into the converter chassis, where the oscillator board is later to be mounted with the aid of spacing pieces. The crystal is now soldered to the mounting plate with the connecting wires facing upwards. A crystal holder was not used because the crystal need not be exchanged, and because contact difficulties should be avoided. The trimmer capacitors and the already wound coil formers are now mounted. The connection points of the crystal, feed-through capacitors and the trimmers offer enough possibilities for mounting the additional components, and for this reason the transistors have not been provided with sockets.

The neutralization inductance Ch 1 of the crystal is soldered directly on to the base of the crystal, which is accessable from above. The resistors R 1, R [illegible], R [illegible], R [illegible] and R 7 as well as the zener diode are all mounted under the chassis and connected to the lower ends of the feed-through capacitors.

The transistor T [illegible] of the 2[illegible] MHz amplifier stage is mounted with its connecting leads facing upwards in a cut-out in the oscillator partition (see Fig. 6). The emitter and screening connections are soldered directly above the cut-out to ground. The ensures that the transistor is firmly held, that the base connection is on one side for the joining to the lowest be soldered on the other side to the trimmer C 9.

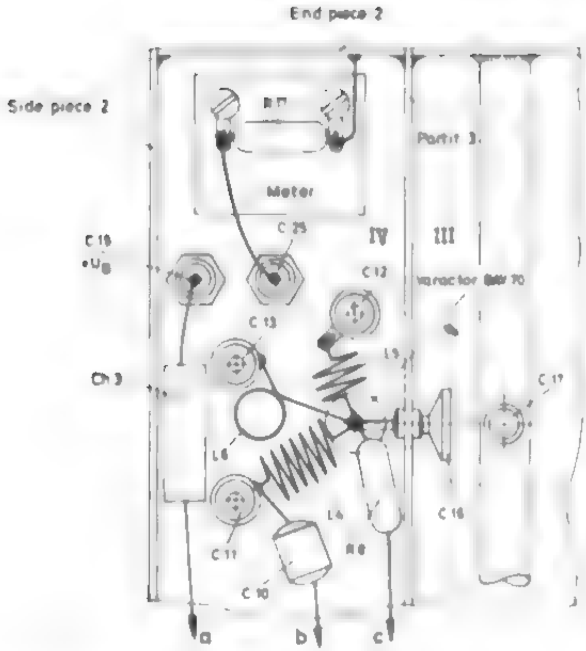
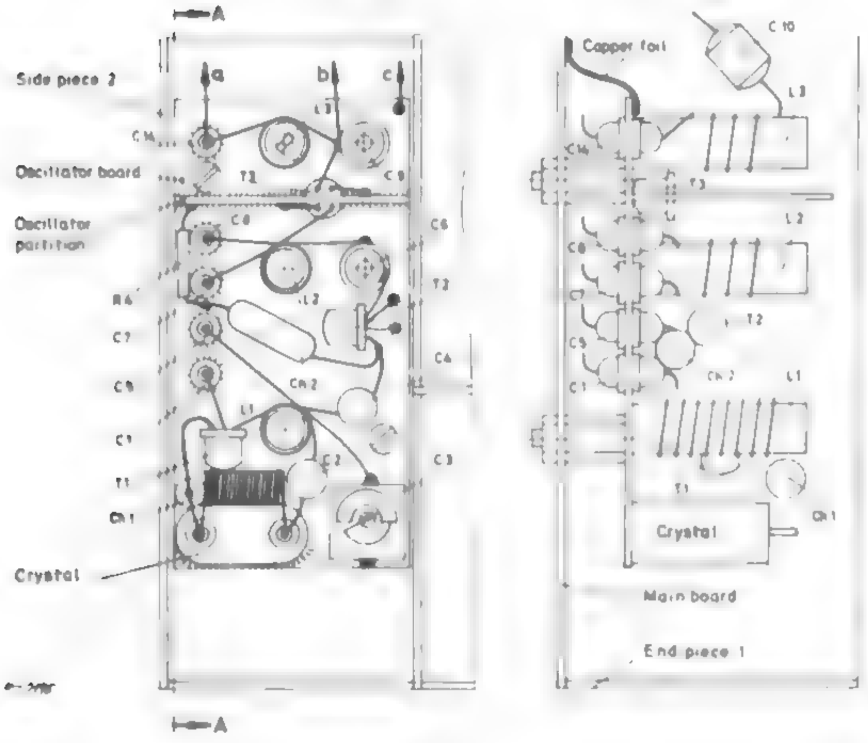


Fig 5:
Mechanical Assembly of the Varactor Multiplier in Chamber IV

DL 3 WR
24cm Converter

Fig 6: (below)
Mechanical Assembly of the Oscillator Portion in Chamber IV, Sectional View A-A (see also Table I)



After mounting the oscillator board into the converter chassis, a 5 mm wide thin copper foil strip is soldered between the upper surface of the oscillator board (at the base of C 9) to the converter chassis (ground point of the varactor diode). The last feed-through capacitor of the oscillator portion C 14 is connected to the external feed-through capacitor C 15 located in front of the meter via a wideband choke. It is pressed into the corner so that the operation of the varactor diode is not disturbed (see Fig. 5). Finally C 10 is soldered between the high point of the 253 MHz circuit and the first varactor trimmer C 11.

3.3. IF PREAMPLIFIER PORTION (Fig. 7 and 8)

This sub-assembly is built up on a printed circuit board which is shorter than chamber V of the converter to enable enough room for the IF feed-through and for the feed-through capacitor C 32. The longest side has a 33 mm by 5 mm cut-out to allow enough room for the plate capacitor C 21 with its two plastic screws holding the mixer diode and for the feed-through capacitor C 25.

The largest components on the printed circuit board are the two 15 by 15 mm large and 26 mm high aluminum screening cans of L 7 and L 8. The arrangement of the components and the printed conductor lanes is shown in Figures 7 and 8.

This sub-assembly is also mounted with two screws soldered to suitable positions and with spacing pieces to the main chassis. The mounting point is sufficient at this frequency for the ground connection and the only connections that are required after assembly is to the mixer diode, i.e. to the plate of capacitor C 21, to C 25, C 32 and to the IF feed-through.

Finally the two feed-through capacitors for the diode current (C 25 and another one not given in the circuit diagram which is to be found before the meter) should be connected to another. Capacitors C 15 and C 32 are also joined together.

A solder tag is fexed to a screw of the IF preamplifier and is used as a ground point. A coaxial cable is now connected to this point and to the IF feed-through. Finally a two core cable is connected to ground and capacitor C 32.

The converter is now fitted with a cover made out of a single coated printed circuit board. The nuts soldered to the corners of the converter and to the corner of the compartment of the primary resonant line, are now used to hold the cover in place. The printed circuit is mounted with the conducting surface facing outwards and grounded only via the five screws. This is sufficient to avoid any detuning of the converter and thus,ensuring that no outside influences effect the operation of the sensitive UHF circuits. If it is desired to make the converter "air-tight", this can be made with copper foil, a rubber padding and a solid iron cover or by sealing all surfaces in a professional manner.

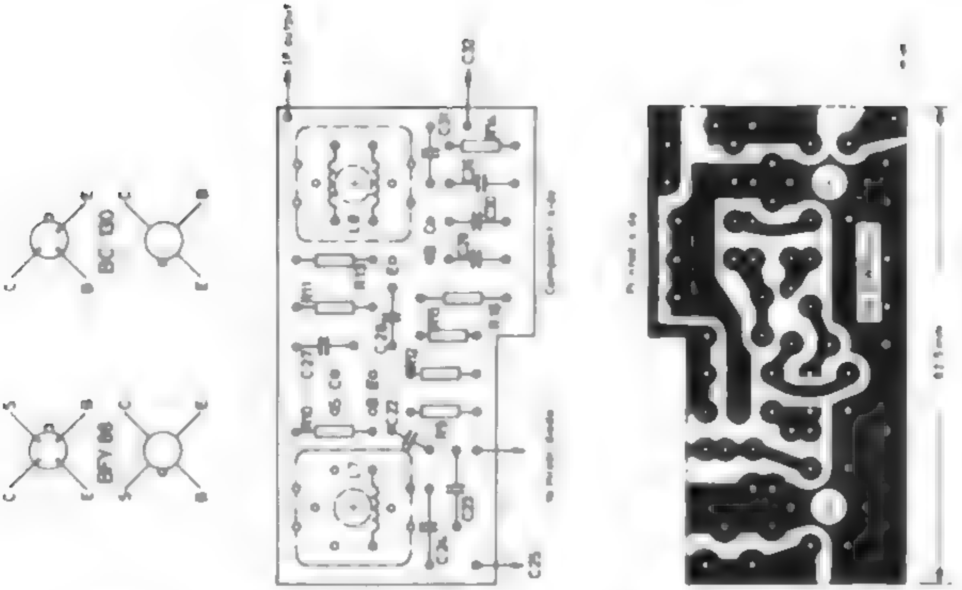


Fig 7 IF Preamplifier of the DL 3 WR 24 cm Converter

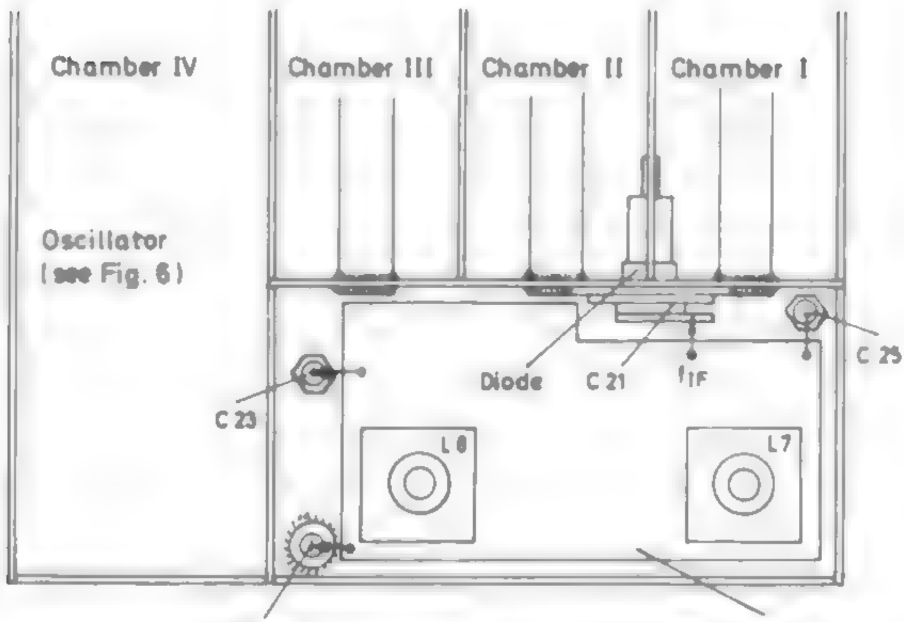


Fig. 8: The Accomodation of the Preamplifier in the 24cm Converter Chassis (DL 3 WR)

3.4. COIL DATA

L 1	9 turns of 1 mm dia. (18 AWG) silver plated copper wire on a 6 mm former without core. Spacing between turns approx. 1 mm. Feed-back and tapping point on the 2nd turn.
L 2	2 turns, wire and former as L 1. Spacing between turns approx. 2.5 mm. Decoupling winding 1 turn of 0.5 mm dia. (24 AWG) enamelled copper wire.
L 3	3.5 turns, wire and former as L 1 except with SW core. Spacing between turns 1 mm.
L 4	6.5 turns of 1 mm dia. (18 AWG) silver plated copper wire wound on a 6 mm former, self-supporting and close wound.
L 5	3 turns of 1 mm dia. (18 AWG) silver plated copper wire wound on a 6 mm former, self-supporting with approx. 1 mm spacing between turns.
L 6	1.5 turns of 1 mm dia. (18 AWG) silver plated copper wire wound on a 6 mm former, self-supporting.
L 7	0.5 μH (approx. 10 turns).
L 8	1 μH (approx. 14 turns) tapping approx. 3 turns.
Ch 1	16 turns of 0.5 mm dia (24 AWG) enamelled copper wire on a 4 mm former, close wound, self supporting.
Ch 2, Ch 3	50 μH wideband choke, uncritical.

4. ALIGNMENT OF THE CONVERTER

The first alignment after adjusting the operating points of transistors T 1 and T 1 is the adjustment of the crystal oscillator. When not oscillating, a collector current of approximately 2 mA will flow. A VHF-receiver detuned to 84 MHz can be used for a coarse frequency check. Attention !! the 7th harmonic of the crystal (125.3 MHz) is in the tuning range of trimmer capacitor C 3. The oscillator should not oscillate at any other frequencies. If this is the case, the neutralization coil should be changed; if the oscillation is at the lower end of the tuning range the number of turns should be increased and vice-versa for oscillations at the higher end.

The alignment is made with the tripler connected, which should pass approximately 3 mA collector current after the alignment when coupled optimumly to the 253 MHz amplifier stage. The coupling to L 1 and the value of R 4 can be varied if necessary. The multiplier stage with transistor T 2 is aligned for the maximum collector current of T 3. T 3 is finally carry out a power measurement, approximately 12 mW should be measured at the third stage.

After installing the above sub-assembly it will be necessary to re-align L 3/C 9, which will have been detuned by the quintupler. This is carried out with the aid of the absorption circuit and by adjusting the core of inductance L 3 until resonance is found. During this alignment, the absorption circuit is taken further and further from the converter.

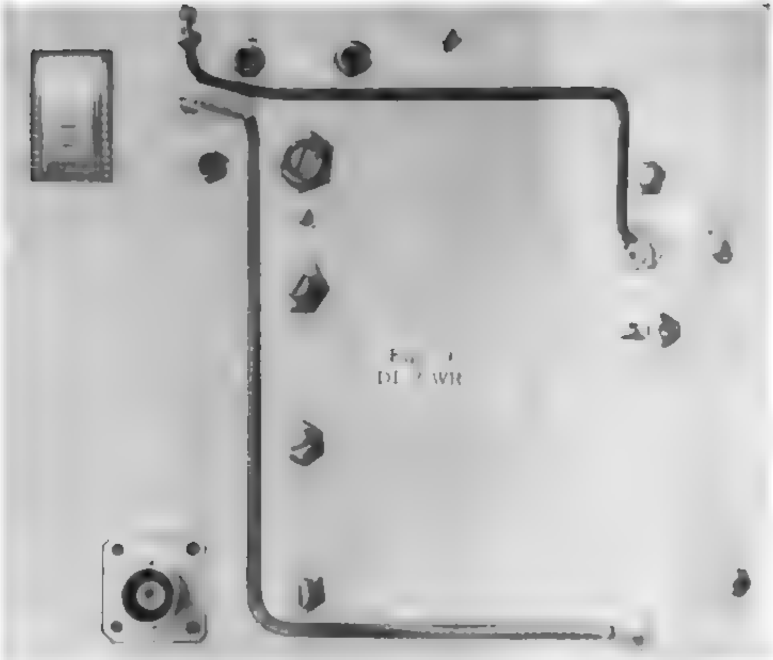
The two trimmers of the UHF resonant line transformer are now carefully aligned, whereby both trimmers should be equally turned (with the prototype converter of the author, the pistons of these two trimmers and that of the input circuit, protruded a little into the holes of the resonant lines). The task of this alignment is to find a position of the trimmers where a diode current of approximately 1 to 2 μA flows. This is measured by disconnecting the built-in meter and by connecting a sensitive μA meter in its place. If the oscillator portion is correctly aligned, the unaligned quintupler will supply enough harmonic power to enable a noticeable small diode current.

The next alignment is the adjustment of the quintupler, whereby the resonance of L 3/C 9 must always be corrected because it is effected by trimmer capacitor C 11. The alignment of the quintupler is correct, if the diode current is at least 1 mA and if no rapid variations of diode current are noticed on carefully rotating trimmers C 11, C 12 and C 13. It is also necessary to readjust the coupling flaps of the primary circuit C 16 and the UHF resonant line transformer during this alignment process. The ideal means of carrying out this alignment is with the aid of a spectrum analyzer which can be simply connected to the input socket of the converter. However, a lecher line can be used to measure the wavelength of the voltages across the idler circuits and the UHF resonant line transformer. The correct wavelength is indicated by a dip in the diode current. The lecher line is either capacitive coupled with the open end or inductively coupled at the closed end. In the first case, the first dip found by varying the position of the short is at a 1/4 wave ($\lambda/4$), whereas with inductive coupling it is at 1/2 wave ($\lambda/4$).

The alignment of the UHF circuits is extremely critical but can be easily carried out with the aid of the M 4 screw thread. If the required diode current is not achieved, the stabilized voltage of the oscillator and quintupler can be modified to 9 or 10 V after the most optimum alignment of all stages has been completed. If on the other hand a too higher diode current is found, this can be corrected by shifting the operating point of the 253 MHz amplifier transistor stage T 3 towards class C operation (R 5 can be removed in this case).

The alignment of the IF amplifier is limited to the checking of the operating points of T 4 and T 5 which should have a collector current of 2 mA and checking for any instability (load input with 500 Ω and the output with 60 Ω). After having correctly aligned the diode current, the IF preamplifier is aligned for maximum noise in the shortwave receiver (10 metres). The alignment of the output stage is clearly heard, whereas the alignment of the first stage of the preamplifier can just be seen on the S meter.

The input circuit of the converter is coarsely tuned by aligning it to the oscillator frequency (dip on the meter) and by making 2 turns in an outwards direction. The resonance point will now be found near this point. The resonance point is found by connecting a correctly terminated antenna or terminating resistor and tuning for the maximum S meter reading (maximum noise). An exact noise alignment and noise matching can only be carried out with the aid of a noise generator (The coupling flap C 20 is approximately 5 mm from the inner conductor).



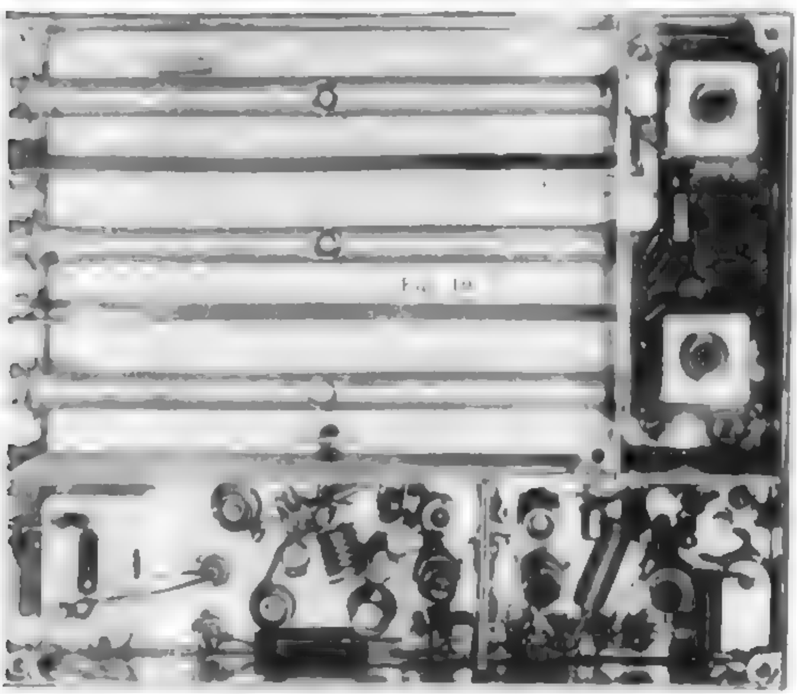


Fig. 10

COMPONENTS LIST (Fig. 2)

RESISTORS

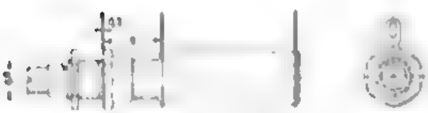
R 1	4.7 kΩ	R 6	3.9 kΩ	R 11, R 13	1.2 kΩ
R 2	220 kΩ	R 7	820 kΩ	R 12	510 kΩ
R 3	12 kΩ	R 8	18 kΩ	R 14	620 kΩ
R 4	56 kΩ	R 9	51 kΩ	R 15	27 kΩ
R 5	100 kΩ	R 10	15 kΩ	R 16	8.2 kΩ
				R 17	100 kΩ

All resistors rated at 0.25 W.

CAPACITORS

C 1, C 5, C 7, C 8, C 14, C 15, C 25, C 32 0.001 µF feed-through capacitor

C 2	820 pF ceramic capacitor (disc or tubular)	C 16	coupling flap, see Fig. 4A/4C
		C 17, C 18, C 19	—
C 3	13 pF trimmer	C 20	coupling flap, see Fig. 3A/4B
C 4	330 pF ceramic capacitor	C 21	approx. 10 pF, see Fig. 4A
C 6	4 pF trimmer	C 22	0.001 µF
C 9	2.5 pF ceramic capacitor	C 23	35 pF
C 10	0.5 pF ceramic capacitor	C 24, C 26, C 28	0.01 µF ceramic
C 11	1 to 10 pF glass	C 27, C 29	0.047 µF capacitor
C 12	0.7 to 5 pF tubular	C 30	27 pF (disc or
C 13	0.5 to 2 pF trimmers	C 31	0.022 µF tubular)



Ceramic tubular trimmers

Screening cans as used in the IF pre-amplifier

SEMICONDUCTORS

T 1 : 2 N 708, 2 N 918
T 2 : BF 115, 2 N 918
T 3 : BFY 88, BFY 90, 2 N 3478
T 4 : BFY 88, BFY 90, 2 N 3478
T 5 : BC 130, BC 108, 2 N 3398, SK 3020

D 1 : Zener 7 V / P_{tot} = 250 mW

D 2 : Varactor : Cj = 5 pF at U_r = 2 V (see text)

D 3 : Mixer diode : 1 N 21 A... F or 1 N 23 ... of Microwave Ass.

OTHER COMPONENTS

1 Crystal: 84.533 MHz HC-6/U with soldering leads
1 Meter: 1 mA, resistance 100 Ω
1 BNC socket UG-1094/U 1 ceramic feed-through

Amount	Figure	Designation, Material, Processing, Surface	Dimensions
1	1	Main board, glass-fibre reinforced epoxyd 1.5 mm board with copper coating on both surfaces	140 x 110 mm
1	-	Cover, glass-fibre reinforced epoxyd 1.5 mm board with single copper coating	140 x 110 mm
1	2	Side piece 1, glass-fibre reinforced epoxyd 1.5 mm board with copper coating on both surfaces	140 x 40 mm
1	2	Side piece 2, material and dimensions as for side piece 1	
1	1	Rear piece 1, glass-fibre reinforced epoxyd 1.5 mm board with copper coating on both surfaces	110 x 40 mm
1	4a	Rear piece 2, 1 mm brass plate	110 x 40 mm
1	3	Partition 1, 0.5 mm brass plate	100 x 40 mm
1	3	Partition 2, 0.5 mm brass plate	100 x 40 mm
1	3	Partition 3, glass-fibre reinforced epoxyd 1.5 mm board with copper coating on both surfaces	140 x 40 mm
1	4A	Partition 4, 0.5 mm brass plate	75 x 40 mm
1	4A, 4D	Capacitor board (C 21), 1 mm brass plate	27 x 17 mm
1	4A, 4D	Foil (dielectric of C 21), 0.2 mm thick teflon	30 x 20 mm
1	4A, 4D	M 3 screws, 10 mm long, nylon	
1	4A, 4D	Contact spring for mixer diode removed from a EL 34 tube socket (Octal)	
3	4A	Inner conductor, 7 mm brass tube, 104 mm long and 0.5 mm wall thickness	
1	4A, 4B	Coupling flap for input 0.5 mm brass plate	
1	4A, 4C	Coupling flap for quintupler, 0.5 mm brass plate	approx. 25 x 10 mm
1	4C, 5	Ceramic solder tag, 8 mm high	
	6	Base board for the oscillator portion, glass-fibre reinforced epoxyd 1.5 mm board with copper coating on both surfaces	70 x 30 mm
1	6	Partition for oscillator portions, same material as the base board	30 x 25 mm
1	7	IF preamplifier board, 1.5 mm pertinax epoxyd board with single copper coating	62 x 24 mm
1	4A, 4D	Leaf spring for mounting mixer diode	
4	-	M 3 screws, 15 mm long with nuts	
5	-	M 3 screws, 8 mm long, the nuts are soldered	

Table 1, List of the required cabinet pieces including accessory parts. (See Fig. 2 to 8)

5. NOTES

An examination of the pass-band curve and an amplification measurement could, unfortunately, not be carried out. However, no fall-off of the noise level can be noticed within the 2 MHz bandwidth. The S meter indication of the noise level however indicates an amplification of approximately 26 dB.

The author would like to thank the amateurs Claus Sondhaus, DJ 2 LI and Hans von Ellen, DJ 1 VO for their kind theoretical and practical assistance in the planning, assembly and alignment of this converter and for the 24 cm QSO's in the vicinity of Ulm, Western Germany.

6. REFERENCES

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A 144 MHz ADAPTER FOR USE WITH A 10 METRE SSB TRANSCEIVER

by E. Krahé, DL 9 GU

The following article describes a complete transverter-converter combination, allowing a SSB transceiver to be operated on the 2 metre band. The described combination is used in conjunction with the 28 MHz range of a Drake TR 4 but is, of course, equally suitable for use with other transceivers having the range 28 to 30 MHz, and for all operating modes (AM, FM and CW).

The adapter is self contained complete with power supply in its own cabinet and is placed between the transceiver and the antenna. The unit also possesses a change-over switch which makes it possible to rapidly switch from VHF to 10 metre operation.

1. THE ELECTRICAL DESIGN

The block diagram of the SSB adapter for 144 MHz is shown in Fig. 1. The circuit diagrams of the individual sub-assemblies are given in Figures 2 to 6. Special attention has been made during the conception of this unit to avoid harmonic and spurious signal generation.

1.1. BLOCK DIAGRAM (Fig. 1)

The operation of the adapter is shown in the block diagram Fig. 1. The auxiliary frequency of 116 MHz is generated in an oscillator comprising tubes V 1, V 2 and V 3 and is fed via output connector 1 to the 2 metre converter (receive mixer) as well as via output connector 2 at a level of 400 mW to the transmit mixer comprising tubes V 4 and V 5. The 28 MHz signal from the transceiver is taken from the control grid of the transceiver PA and fed via a 3 pF capacitor at a level of approximately 1 V to mixer V 4. The 28 MHz signal is now mixed with the auxiliary frequency of 116 MHz, available at connector 4, to obtain the final frequency of 144 MHz, which is then passed via output connector 5 to the linear amplifier V 6. The signal is then amplified by V 6 and fed to the transmit contact of the coaxial change-over relay. The operating voltages for the linear amplifier are obtained from the transceiver, i.e. the plate voltage of + 650 V and the switchable screen grid voltage of + 200 V.
The common auxiliary frequency of 116 MHz is fed to the 2 metre converter via connector 8, after which it is mixed with the 144 MHz signal available at input connector 10. The resulting intermediate frequency of 28 MHz is fed to the transceiver for further processing.

The switching from VHF to 10 metre operation is made with switch S 1. During short wave operation, switch contact S 1a feeds the screen grid voltage back to the transceiver and contact S 1b switches the short wave antenna to the antenna connector of the transceiver. During 2 metre operation, the converted receive signal of 28 MHz is fed via contact S 1b to the receiver and, during transmit, the screen grid voltage from the transceiver is fed to the 2 metre linear amplifier.

The contacts of switch S 1 should be silver plated and be mounted on a ceramic base. This is to ensure that it is able to handle the RF output of the short wave transmitter, which is from 80 to 100 watt in the case in question.

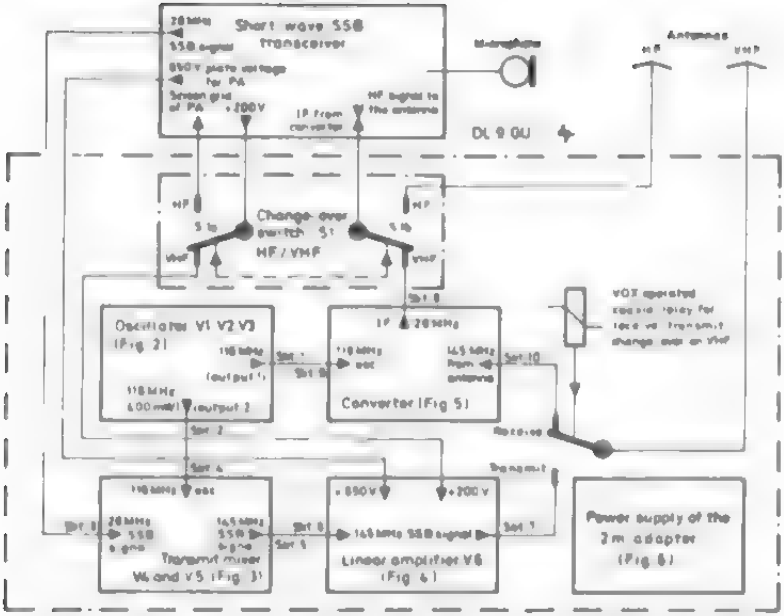


Fig. 1 Block diagram of the adapter to increase the range of a SSB transceiver (10 m) to cover the 2 metre band

1.2. THE INDIVIDUAL CIRCUIT DIAGRAMS OF THE SUB-ASSEMBLIES (Fig. 2 to 6)

The individual circuit diagrams of the sub-assemblies are as follows: Fig. 2 for the oscillator, Fig. 3 for the transmit mixer, Fig. 4 for the linear amplifier, Fig. 5 for the receive converter and Fig. 6 for the power supply.

1.2.1. THE OSCILLATOR (Fig. 2)

The complete circuit diagram of the oscillator stage is shown in Fig. 2. This oscillator supplies the common auxiliary frequencies of 116 MHz for both the receive and transmit mixers. In oscillator V 1, the crystal serves as a cathode coupling link between the two triode systems of the 12 AT 7. The first triode system is a grounded grid amplifier and the second is a cathode-follower for the feed-back. The oscillator circuit - a Butler circuit - is activated at the series resonance of the crystal, which is 58 MHz in the case in question. This circuit has the advantage of being noticeably stabler than traditional circuits using an "inductive" crystal.

The anode circuit of the second system is tuned to the first harmonic (116 MHz). The auxiliary frequency for the receive converter is taken via a capacitor of 2.4 pF from the anode of the 6 J 4 buffer tube V 2 and is fed to the output connector 1. The 116 MHz signal is passed to a further grounded-grid amplifier stage V 3 comprising a second 6 J 4 and is fed to the second output connector 2 at a level of 400 mW for the transmit mixer. The operating voltage V 1, V 2, V 3 and the heater voltages are obtained from the common power supply used for all of the sub-assemblies.

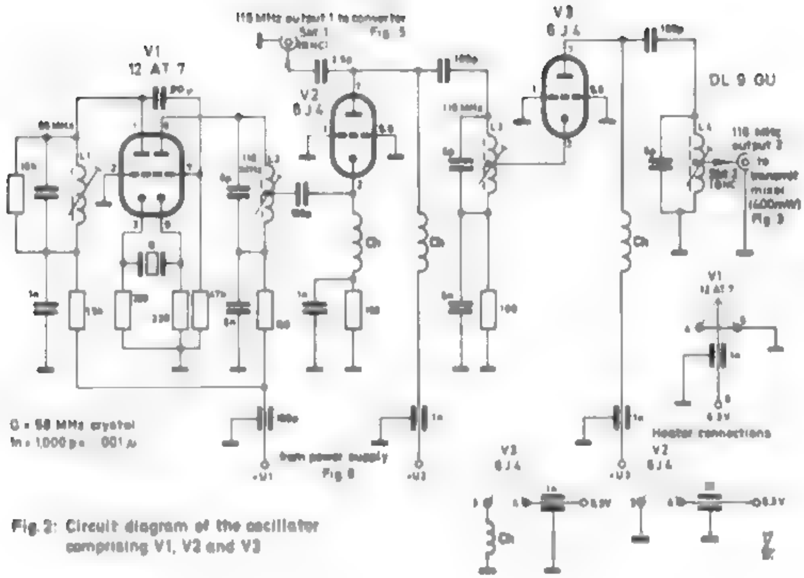


Fig. 2: Circuit diagram of the oscillator comprising V1, V2 and V3

Coil data of the oscillator stage:

L 1	8 turns of 0.6 mm (23 AWG) silver-plated copper wire wound on a 6 mm former. Mounted either on a core-less former or self-supporting.
L 2 and L 3	6 turns of 0.6 mm dia. (23 AWG) silver-plated copper wire wound on a 6 mm dia. former. Mounted either on a core-less former or self-supporting. The tap is made 2 turns from the cold end.
L 4	5 turns of 0.6 mm dia. (23 AWG) silver-plated copper wire wound on a 6 mm dia. former. Mounted either on a core-less former or self-supporting. The tap is made 1.5 turns from the cold end.
Ch	Wideband choke comprising a ferrite bead.

The individual circuits can be aligned without difficulty with the aid of a grid-dip meter.

1.2.2. THE TRANSMIT MIXER (Fig. 3)

The 28 MHz SSB signal is taken via a 3 pF capacitor at an amplitude of approximately 1 V from the control grid of the transceiver PA stage and is fed via a coaxial cable to connector 3 of the transmit mixer. The signal is then fed to the control grid of the mixer tube V 4, an EC 8010. The oscillator frequency available at connector 4 is now fed via a 0.001 µF capacitor to the cathode of the mixer tube. The 144 MHz signal is now available at the anode of the mixer and is fed to the subsequent grounded grid amplifier stage V 5 where it is linearly amplified. The 144 MHz signal is now fed at a power of approximately 2 W to the output connector 5 and thus to the linear amplifier V 6.

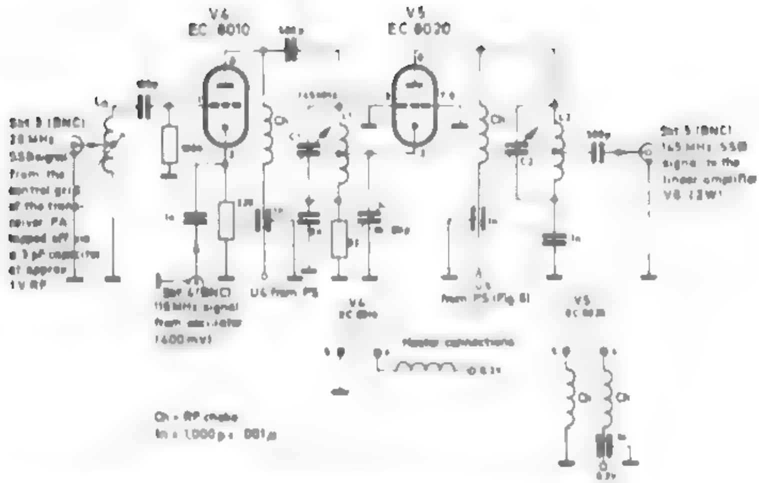


Fig.3 Circuit diagram of the transmit mixer comprising mixer V4 and amplifier V5

A trimmer capacitor of 10 to 30 pF is to be found at the cathode of the EC 8020 (V 5) which is used to match it to the anode circuit of the mixer V 4. A by-pass capacitor of 0,001 μF is connected as close as possible between the cold end of inductance L 1 and ground. The capacitors C 1 and C 2 in the anode circuits of mixer V 4 and tube V 5 are variable capacitors which can be operated from the front panel. The operating voltages U 4, U 5 and the heater voltages are obtained from the common power supply (see Fig. 6).

Coil data for the transmit mixer stage:

L 0 Resonant circuit for 28 MHz

The tap is made 3 turns from the cold end.

L 1 4 turns of 1,5 mm dia. (15 AWG) silver-plated copper wire wound on a 6,5 mm dia. former, self-supporting. The tap is made 1 turn from the cold end.

L 2 3 turns of 1,5 mm dia. (15 AWG) silver-plated copper wire wound on a 10 mm dia. former, self-supporting.
The tap is made 0,5 turns from the cold end.

Ch Wideband choke comprising a ferrite bead.

1.2.3. THE LINEAR POWER AMPLIFIER

The 144 MHz signal is fed via a short coaxial cable to connector 6 of the linear amplifier equipped with the coaxial tetrode 4 X 150 A. The inductances of this stage are in the form of normal coils and the stage offers no special features. The cathode of the 4 X 150 A has multiple connections, each of which is connected via a by-pass capacitor of 4700 pF and a resistor of 12 Ω to the nearest ground point. The capacitor designated Cs in Fig. 4 is the built in by-pass capacitor in the special tube socket of the 4 X 150 A of approximately 2700 pF. The variable capacitor C 2 is

used for the tuning of the anode circuit and can be operated from the front panel. A meter is to be found in the cathode circuit for the monitoring of the cathode current of the tube. The meter measures the voltage across the 4 parallel resistors R x, whereby 600 mV correspond to 200 mA cathode current.

The negative grid bias of the linear amplifier V 6 is obtained from the power supply (see Fig. 6). The plate voltage of + 650 V from the SSB transceiver is permanently fed to the anode of the 4 X 150 A, whereas the screen grid voltage is only fed to the linear amplifier during VHF operation. The VHF output signal from the linear amplifier is fed via connector 7 to the antenna. The power output is 60 W with a plate voltage of 600 V.

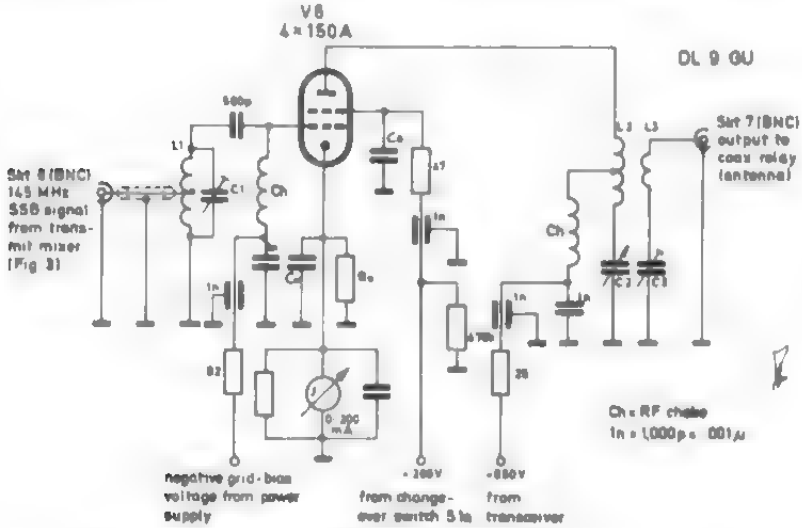


Fig. 4: Circuit diagram of the linear amplifier

Coil data and trimmer capacitors for the linear amplifier stage:

L 1 2 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 6 mm former, self-supporting.

L 2 4 turns of 2 mm dia. (12 AWG) silver-plated copper wire on a 12 mm former, self-supporting with centre tap.

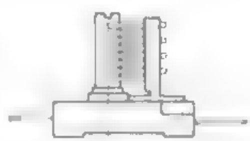
L 3 1 turn of 2 mm dia. (12 AWG) silver-plated copper wire wound on a 10 mm dia. former, self-supporting.

Ch Wideband choke comprising a ferrite bead.

C 1 Trimmer capacitor 2 to 6 pF.

C 2 Variable capacitor 16 pF.

C 3 Trimmer capacitor 50 pF.



1.2.4. THE 2 METRE CONVERTER (Fig. 5a)

The circuit diagram of the 2 metre converter is given in Fig. 5a and the chassis arrangement in Fig. 5b. The converter is equipped with three nu-

vistor tubes 6 CW 4. The 144 MHz signal is fed via connector 10 to the cascode stage comprising V 7 and V 8. The auxiliary frequency of 116 MHz is fed via connector 8 from the oscillator to the mixer stage V 9 where it is mixed with the 144 MHz input signal, thus resulting in the 28 MHz IF signal which is in turn fed via connector 9 to the 10 metre receiver. The connections to the coaxial antenna relay and to the other sub-assemblies can be seen in Fig. 1. The alignment is carried out in the normal manner.

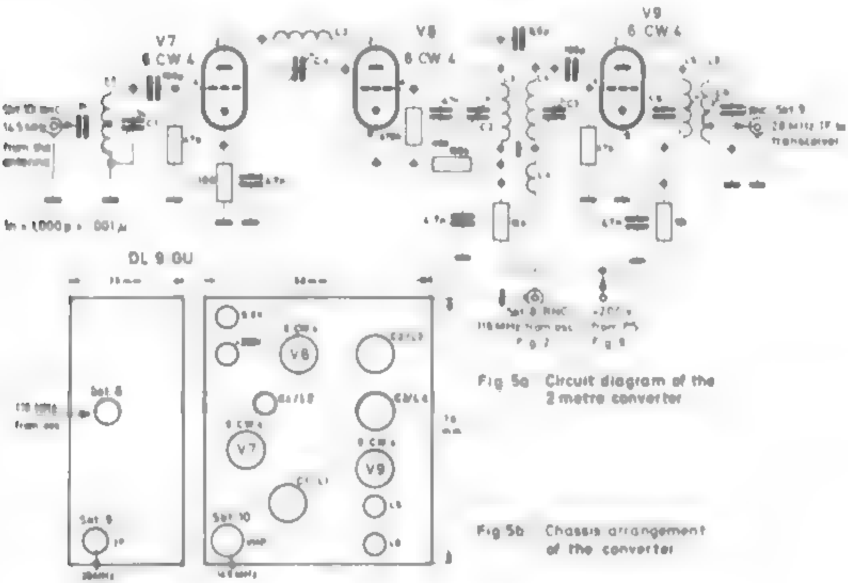


Fig 5a Circuit diagram of the 2 metre converter

Fig 5b Chassis arrangement of the converter

Coil data and trimmer capacitors for the 2 metre converter:

L 1	2.5 turns of 0.6 mm dia. (23 AWG) silver-plated copper wire wound on a 6 mm dia. former. Self-supporting. The tap is made 1 turn from the cold end.
L 2	5 turns of 0.6 mm dia. (23 AWG) silver-plated copper wire wound on a 6 mm dia. former. Self-supporting.
L 3	5 turns; otherwise as L 1
L 4	4 turns; otherwise as L 1
L 4a	2 turns; otherwise as L 1 This coil is built up as a continuation of L 4 and possesses a common axis with L 4.
L 5 and L 6	28 MHz IF transformer Bring the circuits to resonance at 28 MHz with the aid of a grid-dip meter. The tap of L 6 must be found by experiment.
C 1, C 2, C 3 and C 4	1 to 6 pF trimmer capacitors
C 5 and C 6	trimmer capacitors of approx. 10 pF.

1.2.5. COMMON POWER SUPPLY (Fig. 6)

The circuit of the power supply is shown in Fig. 6 and does not contain any special features. The voltages required for the various sub-assemblies are obtained via the corresponding voltage dropper resistors. The common power supply only provides the grid-bias voltage and the heater voltage for the final amplifier, the plate and screen grid voltages are obtained from the short wave transceiver.

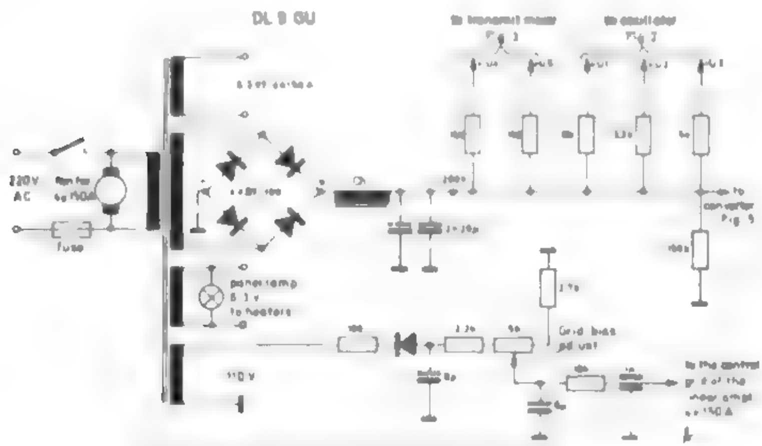


Fig 6 Circuit diagram of the common power supply

2. MECHANICAL ASSEMBLY OF THE 144 MHz TRANSCEIVER ADAPTER

The mechanical build-up of the complete adapter can be seen in the photograph Fig. 7. The dimensions of the complete unit are 265 mm wide, 145 mm high and 185 mm deep. The oscillator portion will be seen as an elongated chamber on the left of the unit with the dimensions of 30 x 40 x 165 mm. The sub-chassis of the transmit mixer is visible at the front of the adapter and the chassis lay-out is shown in Fig. 8. The chassis is made out of brass plate, has the dimensions 30 x 50 x 175 mm and is split into three chambers. Capacitors C 1 and C 2 in the anode circuits of V 4 and V 5 are adjustable from the front panel, which is also true of the anode capacitor C 2 of the linear amplifier tube V 6.

The 2 metre converter is located directly behind the transmit mixer chassis. It will be noticed that the nuvistor converter is extremely small and is built-up on an enclosed chassis. The dimensions and the chassis lay-out are shown in Fig. 5b. The linear amplifier tube is enclosed in a screened compartment to the right of the converter, where the 4 x 150 A can be clearly seen with its ceramic chimney. The anode circuit is connected to the anode radiator with the aid of a metal clamp. The grid circuit of the linear amplifier is to be found underneath the main chassis and is screened with a metal cap which is open in the direction of the fan. A radial fan is used to cool the 4 x 150 A and can be seen at the rear of the linear amplifier compartment. The various sub-assemblies are interconnected with 52 Ω coaxial cables.

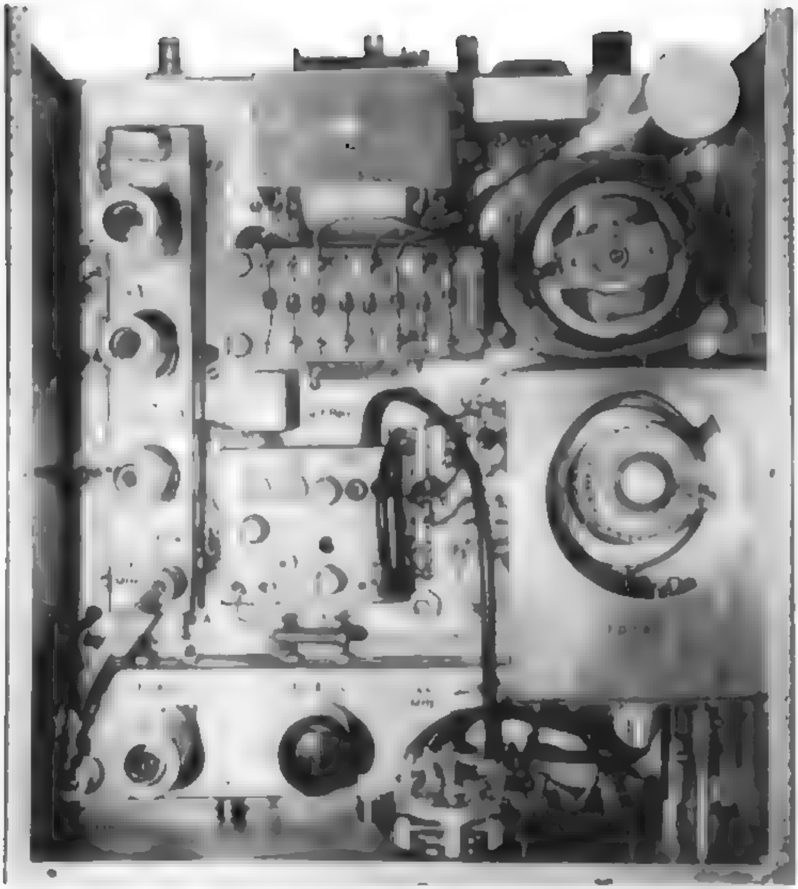


Fig. 7 The 144 MHz adapter for short wave transceivers

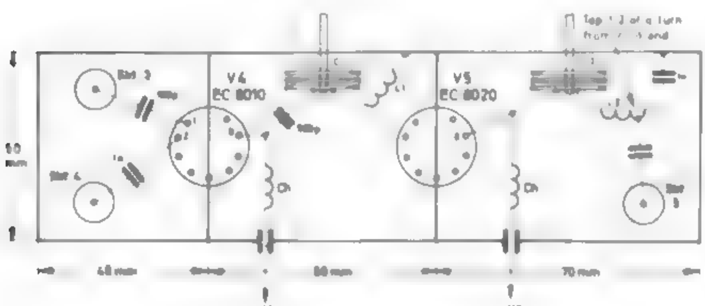


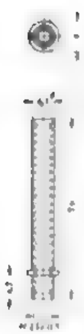
Fig 8 Diagram of the transmit mixer chassis comprising tubes V4 and V5

Chassis 30mm high, from 0.5mm brass plate

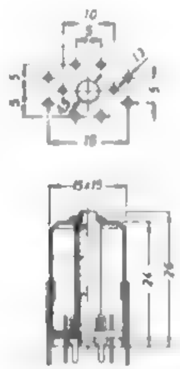
AWG	mm	AWG	mm	AWG	mm
40	0,08	24	0,51	12	2,05
38	0,10	23	0,57	11	2,31
36	0,13	22	0,64	10	2,59
34	0,16	21	0,72	9	2,91
32	0,20	20	0,81	8	3,26
31	0,23	19	0,91	7	3,67
30	0,25	18	1,02	6	4,12
29	0,29	17	1,15	5	4,62
28	0,32	16	1,29	4	5,19
27	0,36	15	1,45	3	5,83
26	0,40	14	1,63	2	6,54
25	0,45	13	1,83	1	7,35

Ceramic sub-micro disc trimmers
as used in DL 6 SW 2 metre converter

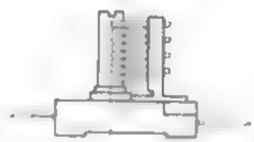
Ceramic tubular trimmers as used in DL 3 WR 24 cm converter



Coil formers
as used in
DL 6 SW
2 metre converter



Screening cans
as used in the IF
pre-amplifier of
the DL 3 WR
24 cm converter



Air spaced trimmers
as used in DL 9 JU
144 MHz/432 MHz
transv.

Centimetre / inches

		0	1	2	3	4	5	6	7	8	9	Io cm
cm	0	0	.394	.787	1.181	1.575	1.969	2.362	2.756	3.150	3.543	3.937
"	.1	.039	.433	.827	1.221	1.614	2.008	2.402	2.795	3.189	3.583	
"	.2	.079	.472	.866	1.260	1.654	2.047	2.441	2.835	3.228	3.622	
"	.3	.118	.512	.906	1.299	1.693	2.087	2.480	2.874	3.268	3.661	
"	.4	.158	.551	.945	1.339	1.732	2.126	2.520	2.913	3.307	3.701	
"	.5	.197	.591	.984	1.378	1.772	2.165	2.559	2.953	3.347	3.740	
"	.6	.236	.630	1.024	1.417	1.811	2.205	2.598	2.992	3.386	3.780	
"	.7	.276	.669	1.063	1.457	1.850	2.244	2.638	3.032	3.425	3.819	
"	.8	.315	.709	1.102	1.496	1.890	2.284	2.677	3.071	3.465	3.858	
"	.9	.354	.748	1.142	1.535	1.929	2.323	2.717	3.110	3.504	3.898	

Millimetre / inches

		0	1	2	3	4	5	6	7	8	9	10 mm
mm	0	0	.0394	.0787	.1181	.1575	.1969	.2362	.2756	.3150	.3543	.3937
"	.1	.0039	.0433	.0827	.1221	.1614	.2008	.2402	.2795	.3189	.3583	
"	.2	.0079	.0472	.0866	.1260	.1654	.2047	.2441	.2835	.3228	.3622	
"	.3	.0118	.0512	.0906	.1299	.1693	.2087	.2480	.2874	.3268	.3661	
"	.4	.0158	.0551	.0945	.1339	.1732	.2126	.2520	.2913	.3307	.3701	
"	.5	.0197	.0591	.0984	.1378	.1772	.2165	.2559	.2953	.3347	.3740	
"	.6	.0236	.0630	.1024	.1417	.1811	.2205	.2598	.2992	.3386	.3780	
"	.7	.0276	.0669	.1063	.1457	.1850	.2244	.2638	.3032	.3425	.3819	
"	.8	.0315	.0709	.1102	.1496	.1890	.2284	.2677	.3071	.3465	.3858	
"	.9	.0354	.0748	.1142	.1535	.1929	.2323	.2717	.3110	.3504	.3898	

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